A GENERALIZED SOLID STATE PROTECTION SCHEME FOR THREE PHASE INDUCTION MOTORS

A Thesis Submitted
in Partial Fulfilment of the Requirements
for the Degree of

MASTER OF TECHNOLOGY

BY

SUNIL KUMAR GARG

to the

DEPARTMENT OF ELECTRICAL ENGINEERING
INDIAN INSTITUTE OF TECHNOLOGY, KANPUR

MAY, 1984

This little work is dedicated to all those who have devoted themselves to explore knowledge to overcome the sufferings of mankind and impart an everlasting smile to their lips.

CERTIFICATE

It is certified that this work "A GENERALIZED SOLID STATE PROTECTION SCHEME FOR THREE PHASE INDUCTION MOTORS" has been carried out under my supervision and has not been submitted elsewhere for a degree.

April, 1984.

(Dr.S.GUPTA)
Professor
Department of Electrical Engineering
Indian Institute of Technology Kanpur

ACKNOWLEDGEMENTS

Words can not suffice to express the gratitude I owe to Dr.S.Gupta, my guide, for suggesting this topic and actively participating to overcome difficulties faced at each level. I am also indebted to him for his encouragement and help he has provided at times when I needed his wise counsel most.

I acknowledge with thanks the authorities of U.P.State Electricity Board their granting me the permission for joining M.Tech. on part-time basis and to undertake the investigation of this problem.

I am grateful to Dr.L.P.Singh and Mr.A.K.Mitra for their active co-operation and help throughout.

I am thankful to Mr.S.N.Sikdar for several occasions of useful discusions regarding this subject and his help in the fabrication and testing of the unit.

Finally, I am obliged to Mr.A.K.Rastogi, Mr.V.P.Tewari, Mr.V.K.Shah, Mr.B.D.Biswas, Mr.Gouri Singh and my wife Kalpana, who shared many responsibilities neglected by me during the course of this work.

-S.K. GARG

INDEX

ABSTRACT

It is observed that Induction Motors are customarily protected with bimetallic relays which are not adequate to safeguard motors under all operating conditions. In the age of electronics even today not much literature has been published in this important field.

Some of the traditional negative sequence networks have been analyzed in detail and improved by making use of the Integrated Circuits and associated devices to get better performance. Induction Motor performance under single phasing with loaded open phase has been carried out in star and Delta connection and verified practically in laboratory to give an idea of various operating parameters for relay setting. A modified current transformer has been suggested for better linearity and reduced cost. For proper caliberation an electronic three phase lab. power supply is designed which can be programmed for simulating different type of voltage unbalance at variable frequencies. A hybrid scheme using voltage and current sensing is designed to achieve multiple functions like motor derating, rate blocking etc. automatically. The emphasis is placed on optimizing the use of each component to make the whole scheme commercially viable.

INTRODUCTION

Induction motors which are termed the work horses of Industry play such a vital role in every process that it becomes essential to protect them against many operating conditions which might arise during the course of it's use. Protection of Induction motors is essential also for the reason that although they are cheap compared to other types of motors, still they cost a substantial amount and the time required to repair any damage is comparable to that required for other motors. This might adversely effect the overall economy of a process in the event of failure or damage of the motor. To avoid such a situation, it is always better to have a protection scheme which can efficiently sense an undesired operating condition and trip the motor preventing long term damage, although the percentage cost of protection vis a vis motor cost may be a little higher compared to this ratio for other type of motors i.e. synchronous motors, D.C. motors etc.

General practice has been to use bimetallic relays which are, beyond doubt very robust for industrial applications, but they are current operated devices and hence they can not discrimate any quantity which is voltage dependent. They discriminate unwanted voltage conditions indirectly by sensing current which is governed by parameters of the equipment on which this voltage is impressed. These bimetallic relays can act as over current sensors and thus prevent total burnout of the equipment but they give very little flexibility where protection strategy is to be based

on several input quantities such as voltage,current,power factor etc etc. Even when operating as overcurrent element, bimetallic relays can not distinguish between a constant overload and a transient overload. Any effort to provide such discrimination ultimately boils down to adding time delay in operation which in turn may be dangerous for the motor. Moreover the performance of these relays is apt to be effected by ambient temperature and humidity also and it becomes difficult to set them accurately.

A simple solid state relay with very limited number of components can provide adequate protection against all the fault conditions and at the same time can be made much faster. The same unit can be used for a wide range of motors by making adjustments in the same unit.

Several single phasing preventers are available in the market using electronic components but they have several limitations. An effort is made to design a protection scheme which protects the motor under different unwanted input conditions keeping in view the fact that unit does not become unnecessarily complicated and costly but still maintaining the functions it is supposed to perform.

The design is based on the requirements of the motors to be protected as mentioned below. The use of electronic devices has been optimized to reduce cost and size. The effort has also been made to make it a generalized design for all kinds of motors. For proper calibration of the unit a lab-test three phase supply is electronically generated which

has the flexibility of simulating different kind of fault conditions which the relay has to sense. This supply eliminates the need to resort to three phase power supply available from main lines as it may not be balanced at the time of testing of the relay. Also the frequency of mains may not be exact 50 cycles. The electronically generated supply can be varied in frequency for doing frequency response test of the relay without effecting the phase and magnitude of the generated voltage.

<u>The need for protection arises out of the following causes</u> :

(1) <u>Due to phase failure, phase sequence change and phase unbalance</u> :

When one phase of a three phase motor fails, it draws large current from rest of the two phases which may be within the operating current range of the motor if it is not heavily loaded. This might damage the insultation of one of the windings of delta connected motor. When phase unbalance occurs, the rotor heating increases greatly although the torque is not greatly effected. With 10% negative seq.voltage, the copper losses increase from 2.8% to 4.5% while torque reduces by .4%[1]. To keep the heating in machine constant the machine should be derated. The derating factor comes out to be 2.2% derating of current for 1% N.S voltage. This dynamic setting is incorporated in the IDMT generator of the unit described in Chapter 5. Phase sequence change may not be desirable for pumps, printing presses, fans etc.

A Negative Sequence filter is designed with proper frequency compensation to derive true Negative Sequence voltage under above mentioned conditions. A detailed analysis of voltage generated in the open phase of a motor running at almost synchronous speed is done in Chapter Two and logically a 5% imbalance condition is derived as optimum for setting of N.S. voltage to trip the motor. Obviously these functions can not be acheived by bimetallic relays.

(2) Due to overload, load jamming and bearing failare :

When the motor is overloaded, heating in winding increases and to avoid insulation failure the motor should trip in a finite time depending upon the amount of overload. Also for short term overload the motor should not trip unnecessarily. An Inverse Definite minimum time stage is used in conjunction with overload sensor to trip the relay. A modified type of CT is also suggested for sensing the current for this stage. A derating feedback from Negative Sequence stage is compounded with this stage.

(3) Due to overvoltage and undervoltage conditions :

During high voltage condition the magnetic path may get saturated and harmonics in the motor may be generated which may cause heating. At low voltage the motor may come to a stand still or may not start at all.

The current during high voltage may not be very large to cause tripping on overloading but harmonics content may be come large to cause overheating hence

overvoltage sensing is a better resort than overcurrent under such condition.

(4) Due to long acceleration, High ON/OFF duty cycle and rapid reversal :

During each of these conditions the motor draws a large current and the motor may be rated for such intermittent operation such that steady state current is still comparable to some other motor of the same rating. With bimetallic relays we may have to set it for larger current hence protection will not be provided under steady state conditions. This section finds out the rate at which the current is falling and if it is more than the adjusted setting the overload section is blocked. During such overloads the low voltage section is also shut off to avoid false tripping of the motor.

(5) Due to loss of load :

Sometimes loss of load may not be desirable for pumps etc. when they suck in air. Such a condition can be sensed and motor stopped by electronic relay.

(6) Overfrequency and underfrequency sensing :

Overfrequency and underfrequency tripping is required in some applications where vibration increase on either side of 50 C/s. Besides this certain gear coupled drives and frictional loads will cause excessive power input at overfrequency beyond

a limit. This sensing is provided in the frequency compensation circuit of the Negative Sequence filter and can be adjusted according to requirement.

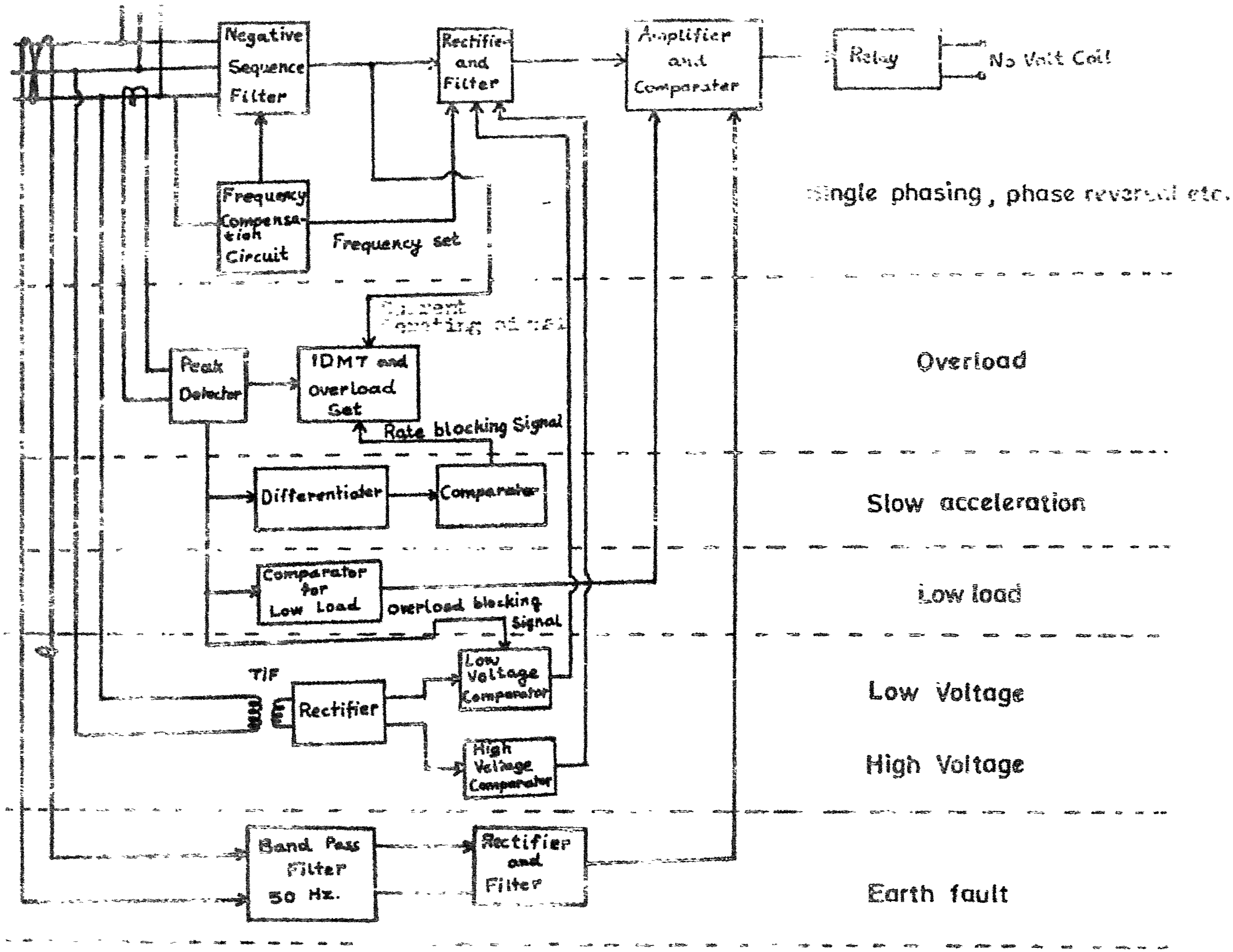
(7) Due to insulation failure of one or more phases :

Ground faults may be developed due to insulation failure which have to be detected as quickly as possible to avoid further damage to the winding and housing. A zero sequence CT surrounding all the three windings is used in conjuction with a 50 HZ band pass filter. The trip current is adjusted to 100 m Amp.

(8) Protection fail/safe arrangement :

This is a common feature of any protection scheme to safeguard the equipment i.e. trip it when input power to protection fail . This is achieved by using the N/O contact to activate the main contactor.

The block diagram of the overall scheme is given on the next page.



Block Diagram of the Overall Scheme

CHAPTER -II

Analysis of Induction Motor Understeady State When One of the Fuses Blows :

Assume a three phase induction motor running at full load having a large inertia rotor. The motors generally run at 2 to 3% slip. Suppose now one of the phase fuses suddenly blows off, due to the inertia of rotor the machine will take some time before the speed reduces. If we neglect the electrical transients in the motor winding, which die out in the order of a few tens of milliseconds, we may take the steady state model of the machines with slowly varying slip. This is important because when one of the fuse blows, the induced EMF in the open phase due to the rotating magnetic field inside the machine may be appreciable and of such polarity and phase which may prevent the tripping of motor, because N.S. circuit is not able to detect the disconnection of motor due to this induced EMF. We first develop the mathematical equations governing this induced EMF assuming absence of electrical transients within the machine. We first analyse a star connected motor because- -motorsare started in star mode.

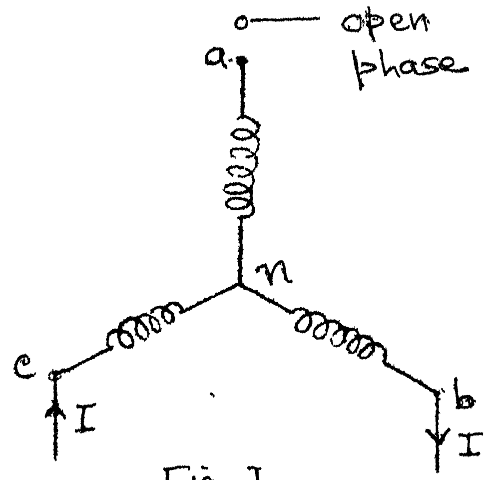


Fig -I

Say phase 'a' is opened out in figure 1.

then, $I_b + I_c = 0$, $I_a = 0$

(neutral is not connected)

Therefore $I_b = -I_c = I$

Treating phase 'a' as reference

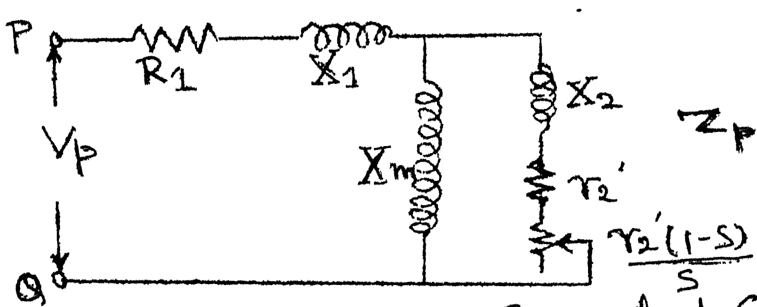
$$I_o = \frac{I_a+I_b+I_c}{3} = 0$$

$$I_p = \frac{I_a+\alpha I_b+\alpha^2 I_c}{3} = \frac{(\alpha-\alpha^2)I}{3} = j\,I/\sqrt{3}$$

$$I_n = \frac{I_a+\alpha^2 I_b+\alpha I_c}{3} = -j\,I/\sqrt{3}$$

Where I_p and I_n are positive and negative sequence currents respectively.

Now from two revolving field theory of induction motor we take the positive sequence and negative sequence equivalent circuit which do not interact on each other due to the symmetry of the induction machine structure and balanced three phases.

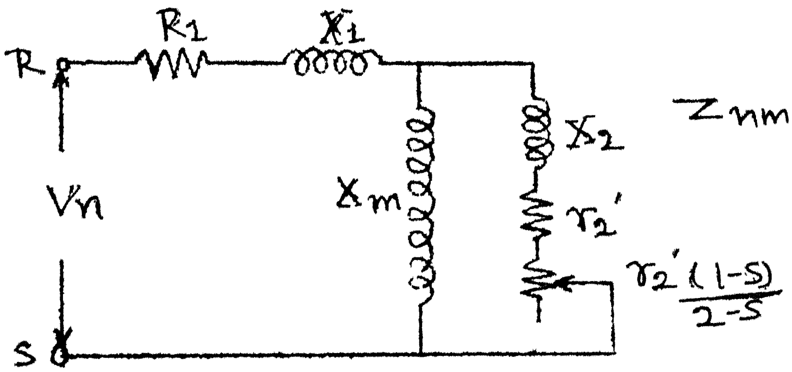


$$Z_{pm}(s) = (R_1 + jX_1) + jX_m // \left(\frac{r_2'}{s} + jX_2\right)$$

Positive Sequence Equivalent Circuit

. Figure 2.

In fig 2. Between terminals P and Q is shown the positive sequence impedance of the motor. The circuit I_p develops a positive sequence voltage V_p across PQ which does not interact with negative sequence network which is shown below.



$$Z_{nm}(s) = (R_1 + jX_1) + jX_m // \left(\frac{r_2'}{2-S} + jX\right.$$

Negative Seq. Equivalent Ckt.
Fig-3.

In fig 3. Between terminals R and S is shown the negative sequence equivalent of induction motor. Current I_n produces negative seq. voltage V_n.

Treating the positive seq. impedance = $Z_{pm}(s)$ and Negative seq. impedance as $Z_{nm}(s)$ where 's' refer to slip and not the 's' domain, we get

$V_p = I_p Z_{pm}$ and $V_n = I_n Z_{nm}$, $V_z = 0$ (zero seqs) voltage

Now $\quad V_{cn} = \alpha V_p + \alpha^2 V_n + V_z$ $\quad$ in fig I.

$\quad\quad V_{bn} = \alpha^2 V_p + \alpha V_n + V_z$

Therefore, $V_{cb} = V_{cn} - V_{bn} = (\alpha - \alpha^2)\, V_p + (\alpha^2 - \alpha)\, V_n$

$$= j\sqrt{3}\, V_p - j\sqrt{3}\, V_n$$

or $\quad V_p - V_n = \dfrac{V_{cb}}{j\sqrt{3}}$

Therefore, $I_p Z_{pn} - I_n Z_{nm} = \dfrac{V_{cb}}{j\sqrt{3}}$

Therefore, $j\dfrac{I}{\sqrt{3}}\, Z_{pm} + j\dfrac{I}{\sqrt{3}}\, Z_{nm} = j\dfrac{V_{cb}}{\sqrt{3}}$

or $\quad I = \dfrac{-V_{cb}}{Z_{pm} + Z_{nm}} \quad \ldots\ldots \text{(I)}$

Therefore, $V_{an} = V_p + V_n + V_z = I_p Z_{pm} + I_n Z_{nm} + 0$

$$= j\frac{I}{\sqrt{3}}\,(Z_{pm} - Z_{nm})$$

Now from I

$V_{an} = -j\dfrac{V_{cb}}{\sqrt{3}}\; \dfrac{(Z_{pm} - Z_{nm})}{Z_{pm} + Z_{nm}}$ $\quad$ where Z_{pm} and Z_{nm} are functions of slip.

Now looking at the phasor diagram in fig 4.

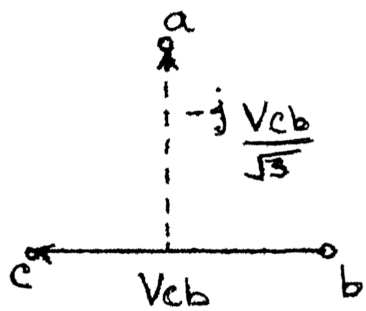


Fig 4.

V_{an} is perpendicular to V_{cb} and further rotated by

$$\frac{Z_{pm} - Z_{nm}}{Z_{pm} + Z_{nm}} = Z\underline{/\theta}.$$ Vector $-j\frac{V_{cb}}{\sqrt{3}}$ is $= V_{an}/\text{normal}$ i.e. when the three phases are healthy and balanced $-j\frac{V_{cb}}{\sqrt{3}}$ is the vector V'_{an} under healthy condition. shown in fig

So, $V_{an} = V_{an}' \dfrac{Z_{pm} - Z_{nm}}{Z_{pm} + Z_{nm}}$ where V_{an}' stands for line to neutral voltage under healthy balanced conditions.

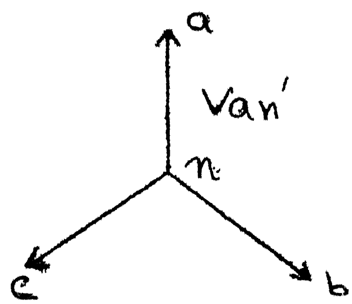


Fig 5.

V_{an} is generated due to the rotating field inside the machine and is a function of positive and negative sequence impedances of the machine which in turn are functions of slip.

Therefore, $V_{an}(s) = V_{an}' \left[\dfrac{Z_{pm}(s) - Z_{nm}(s)}{Z_{pm}(s) + Z_{nm}(s)} \right]$ II

Taking a case of 50 HP motor [2] we have $r_1 = .19\,\Omega/2$, $x_1 = j\ 1.12/2$, $r_2 = \frac{.29\,\Omega}{2}$, $x_2 = j\ 1.12/2$, $x_m = j\ 16.8/2$. At standstill $s = 1$ and $Z_{pm} = Z_{nm}$ so $V_{an}(0) = 0$, so the N.S. network will sense it and trip.

At full load $s = .026$

$r_2'/s = 5.6\Omega$ and $R_2'/2-s = .145\,\Omega$, $\frac{x_2'}{2} = j\,.56\Omega$

Therefore, $Z_{pm} = (.095+.56j) + (5.6 + j\,.56)'' \; j\,8.4$

and $Z_{nm} = (.095 + .56\,j) + (.145 + .56\,j)'' \; j\,8.4$

this shows that at s = .026, $Z_{pm} >> Z_{nm}$

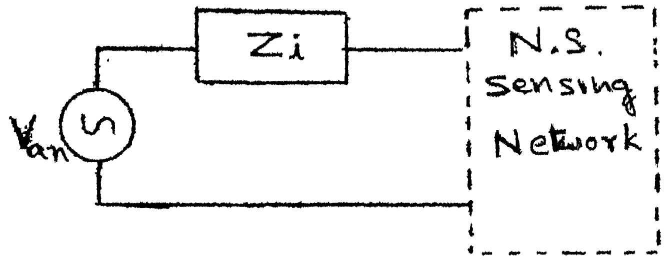
Therefore, from equation II we get

$$V_{an}(s) = V_{an}' \frac{(1-\frac{Z_{nm}}{Z_{pm}})}{(1+\frac{Z_{nm}}{Z_{pm}})} \quad Z_{pm} >> Z_{nm} \simeq V_{an}'$$

Thus the generated voltage will be approximately same as the normal line voltage. Hence a large motor running at full load with large inertia, when suddenly one fuse blows, will generate open circuit voltage of same polarity and phase as the healthy line voltage because the normal slip does not reduce due to the large inertia of motor.

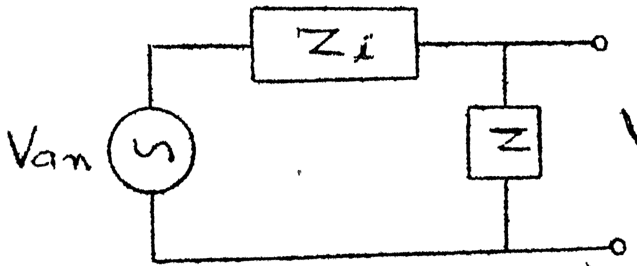
To overcome this difficulty we try a modified N.S. network. Three impedances 'Z' are connected across each phase. The use is made of the fact that as long as all the three phases are alive, they act asideal voltage source and impedance 'Z' acts as a load. When any of the phase is disconnected, impedance 'Z' loads that phase. Assuming that N.S. sensing network is a high impedance network, it acts as an almost open circuit and whatever internal voltage is generated inside the phase, appears at terminals, but if 'Z' is comparatively low impedance, it loads that phase

and causes voltage drop when the fuse blows from that phase.



$V_o = V_{an}$

Equivalent Ckt of open phase without external impedance.
Fig 6.



$V_o < V_{an}$

Equivalent Ckt with external impedance Z.
Fig 7.

<u>Analysis of motor with external impedances Z</u> :

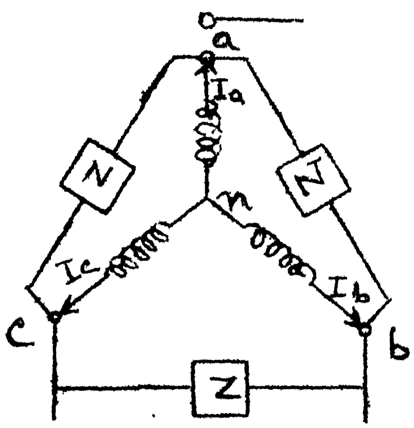


Fig 8.

Fig 8 Phase 'b' and phase 'c' are healthy phases and phase 'a' opens out. Impedance 'Z' is connected across voltage sources V_{cb} so it does not play any role. Rest two impedances 'Z' are connected to phase 'a' which has some internal impedance.

Now,

$$I_a = \frac{V_{an} - V_{bn}}{Z} + \frac{V_{an} - V_{cn}}{Z} = \frac{2V_{an}}{Z} - \frac{(V_{bn} + V_{cn})}{Z} \quad \ldots III$$

also $I_a + I_b + I_c = 0$ so $I_o = 0$

Hence $V_o = 0$

Now denoting positive and negative sequence currents as I_p and I_n.

$$V_p = I_p Z_{pm} \quad \text{and} \quad V_n = I_n Z_{nm}$$

Also from previous analysis

$$V_p - V_n = \frac{V_{cb}}{j\sqrt{3}} \quad \ldots\ldots IV$$

From III,

$$I_p + I_n = \frac{2(V_p+V_n)-(\alpha^2 V_p+\alpha V_n+\alpha V_p+\alpha^2 V_n)}{Z}$$

$$= \frac{2(V_p+V_n)+(V_p+V_n)}{Z} = \frac{3(V_p+V_n)}{Z}$$

or $$\frac{V_p}{Z_{pm}} + \frac{V_n}{Z_{nm}} = \frac{3(V_p+V_n)}{Z}$$

or $$V_p\left(\frac{1}{Z_{pm}} - \frac{3}{Z}\right) = V_n\left(\frac{3}{Z} - \frac{1}{Z_{nm}}\right)$$

or $$V_n = V_p \frac{Z - 3Z_{pm}}{3Z_{nm} - Z} \cdot \frac{Z_{pm}}{Z_{nm}}$$

$$= -V_p \frac{3Z_{pm}-Z}{3Z_{nm} - Z} \cdot \frac{Z_{nm}}{Z_{pm}} \quad \ldots\ldots V$$

Substituting in IV

$$V_p \left(1 + \frac{3Z_{pm} - Z}{3Z_{nm} - Z} \cdot \frac{Z_{nm}}{Z_{pm}}\right) = \frac{V_{cb}}{j\sqrt{3}}$$

$$V_p = \frac{V_{cb}}{j\sqrt{3}\left[1 + \frac{3Z_{bm} - Z}{3Z_{nm} - Z} \cdot \frac{Z_{nm}}{Z_{pm}}\right]} \qquad \ldots. \ VI$$

also, $V_{an} = V_p + V_n$

From V, substituting V_n

$$V_{an} = V_p \left(1 - \frac{3Z_{pm} - Z}{3Z_{nm} - Z} \cdot \frac{Z_{nm}}{Z_{pm}}\right)$$

Substituting the value of V_p from VI

$$V_{an} = \frac{V_{cb}}{j\sqrt{3}} \quad \frac{1 - \frac{3Z_{pm} - Z}{3Z_{nm} - Z} \cdot \frac{Z_{nm}}{Z_{pm}}}{1 + \frac{3Z_{pm} - Z}{3Z_{nm} - Z} \cdot \frac{Z_{nm}}{Z_{pm}}}$$

$$= \frac{V_{cb}}{j\sqrt{3}} \quad \frac{3Z_{pm}Z_{nm} - ZZ_{pm} = 3Z_{pm}Z_{nm} + ZZ_{nm}}{3Z_{pm}Z_{nm} - ZZ_{pm} + 3Z_{pm}Z_{nm} - ZZ_{nm}}$$

$$= \frac{V_{cb}}{j\sqrt{3}} \quad \frac{(Z_{nm} - Z_{pm})\,Z}{6Z_{pm}Z_{nm} - Z(Z_{pm} + Z_{nm})}$$

$$= \frac{V_{cb}}{j\sqrt{3}} \quad \frac{(Z_{pm} - Z_{nm})}{(Z_{pm} + Z_{nm}) - 6Z_{pm}Z_{nm}/Z} \qquad \ldots. \ VII$$

where Z_{pm} and Z_{nm} are functions of slip, while 'Z' is pure impedance independent of slip. As speed reduces Z_{pm} reduces and Znm increases so the voltage generated drops.

If $Z = \infty$ weget

$$V_{an'\, Z = \infty} = \frac{V_{cb}}{j\sqrt{3}} \left[\frac{Z_{pm} - Z_{nm}}{Z_{pm} + Z_{nm}} \right] \text{ same as II}$$

Now say the machine is running at '0' slip so that $Z_{pm} >> Z_{nm}$. Substituting in VII weget.

$$V_{an} = \frac{V_{cb}}{j\sqrt{3}} \quad \frac{Z}{Z - 6Z_{nm}} = V_{an}' \left[\frac{Z'}{Z - 6Z_{nm}} \right] \quad \ldots\ldots \text{ VIII}$$

If 'Z' is now properly chosen, we get V_{an} less than V_{an}' even at zero slip when one of the phase opens out.

From the 50 HP motor data of previous section we obtain $Z_{nm} \simeq (.145 + .56j)$

Suppose we chose $Z = 140\ Z_{nm}$ then from VIII

$$V_{an} = V_{an}' \ \frac{140\ Z_{nm}}{134\ Z_{nm}} = 1.045\ V_{an}'$$

i.e. almost 5% disbalance at zero slip.

Now $140\ Z_{nm} = (20.3 + 78.4j)$ which is very small impedance and causes considerable power loss and loading of each phase under healthy condition of line (500 Watt each phase). It a highly resistive load is applied i.e. say 60 Watt resistors/400 Volt., the Z is so large compared to $6\ Z_{nm}$ that it has hardly any effect on N.S. voltage as has been experimentally verified for a star connected motor.

<u>Experiment</u> :- <u>Motor specifications</u> :- 3 HP, star connected, 400/440 Volts, RPM 1420 .

<u>Observations</u> :- Line to line voltage = 300 Volts

No load current = .5 Amp. Healthy condition

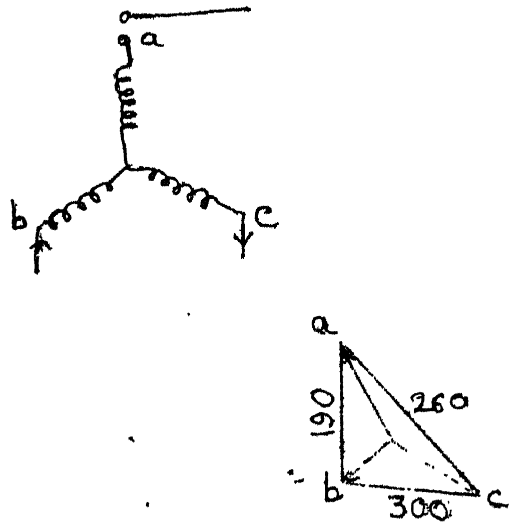


Fig 9.

<u>Phase 'a' opened</u> :-

V_{cb} = 300 Volts.

V_{ac} = 260 Volts.

V_{ab} = 196 Volts.

<u>With 50 Watt load on each phase and phase 'a' opened</u> :

V_{cb} = 300 Volts.

V_{ac} = 260 Volts.

V_{ab} = 188 Volts.

The V_{an} vector bends due to the fact that in star connected mode the positive and negative seq. impedances are high compared to Δ connection and hence $Z_{pm} - Z_{nm}/Z_{pm} + Z_{nm}$ gives appreciable ratio to effect V_{an}' substantially while presence of external 'Z' does not have marked effect on V_{an}.

<u>Analysis of Δ connected motor</u> :-

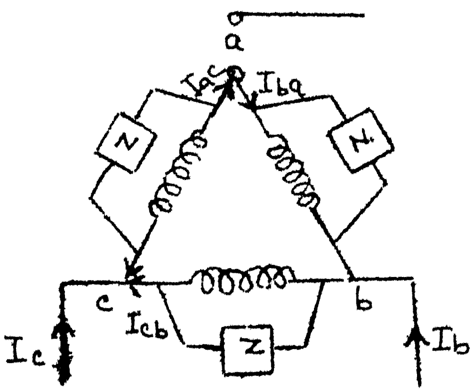


Fig 10

$\underline{I_{ac} = I_{ba}}$ including current through Z , Since $I_a = 0$

$I_{ac} = \alpha^2 I_p + \alpha I_n$ — I   Since I_o and V_o are zero

$I_{ba} = \alpha I_p + \alpha^2 I_n$ — II

$I_{cb} = I_p + I_n$

also,

$$V_{ca} = \alpha^2 V_p + \alpha V_n$$

$$V_{bc} = V_p + V_n$$

$$V_{ab} = \alpha V_p + \alpha^2 V_n$$

From I and II

$$\alpha^2 I_p + \alpha I_n = \alpha I_p + \alpha^2 I_n$$

$$(\alpha^2 - \alpha) I_p = (\alpha^2 - \alpha) I_n$$

or $I_p = I_n$

Now,

$$V_{bc} = V_p + V_n = I_p(Z_p + Z_n)$$

or $I_p = \dfrac{V_{bc}}{Z_p + Z_n}$

where $Z_p = Z_{pm} '' Z$

and $Z_n = Z_{nm} '' Z$

where Z_{pm} and Z_{nm} are positive and negative sequence impedeances of the motor and 'Z' is some external loading impedance.

Now $I_{ac} = \alpha^2 I_p + \alpha I_n$

$$= \frac{(\alpha^2 + \alpha) V_{bc}}{Z_p + Z_n} = \frac{-V_{bc}}{Z_p + Z_n}$$

$$I_{ba} = I_{ac}$$

$$I_{cb} = \frac{2V_{bc}}{Z_p + Z_n}$$

Therefore, $I_c = I_{ac} - I_{cb}$

$$= \frac{-3V_{bc}}{Z_p + Z_n}$$

and $I_b = I_{cb} - I_{ba}$

$$= \frac{3V_{bc}}{Z_p + Z_n}$$

Now $V_{ca} = \dfrac{(\alpha^2 Z_p + \alpha Z_n)V_{bc}}{Z_p + Z_n}$ — III

and $V_{ab} = \dfrac{(\alpha Z_p + \alpha^2 Z_n)V_{bc}}{Z_p + Z_n}$ — IV

From (3)

$$V_{ca} = \frac{\dfrac{\alpha^2 Z_{pm} Z}{Z_{pm} + Z} + \dfrac{\alpha Z_{nm} Z}{Z_{nm} + Z}}{\dfrac{Z_{pm} + Z}{Z_{pm} + Z} + \dfrac{Z_{nm} Z}{Z_{nm} + Z}} \cdot V_{be}$$

$$= \frac{\alpha^2 Z_{pm} Z\,(Z_{nm} + Z) + \alpha Z_{nm} Z\,(Z_{pm} + Z)}{Z_{pm} Z\,(Z_{nm} + Z) + Z_{nm} Z (Z_{pm} + Z)}\, V_{bc}$$

$$= \frac{\alpha^2 Z_{pm} Z_{nm} + \alpha^2 Z_{pm} Z + \alpha Z_{nm} Z_{pm} + \alpha Z_{nm} Z}{2Z_{pm} Z_{nm} + Z(Z_{pm} + Z_{nm})}\, V_{bc}$$

$$= \frac{-Z_{pm} Z_{nm} + Z(\alpha^2 Z_{pm} + \alpha Z_{nm})}{2Z_{pm} Z_{nm} + Z(Z_{pm} + Z_{nm})}\, V_{bc}$$

$$= \frac{Z(\alpha^2 Z_{pm} + \alpha Z_{nm}) - Z_{pm} Z_{nm}}{2Z_{pm} Z_{nm} + Z(Z_{pm} + Z_{nm})}\, V_{bc} \qquad \text{— V}$$

It $Z = \infty$

$$V_{ca} = \frac{\alpha^2 Z_{pm} + \alpha Z_{nm}}{Z_{pm} + Z_{nm}} \quad \text{same as equation III}$$

Rearnanging V

$$V_{ca} = \frac{(\alpha^2 Z_{pm} + \alpha Z_{nm}) - \frac{Z_{pm} Z_{nm}}{Z}}{(Z_{pm} + Z_{nm}) + \frac{2 Z_{pm} Z_{nm}}{Z}} V_{bc}$$

This shows that if we reduce Z to make $\frac{Z_{pm} Z_{nm}}{Z}$ large we get more disbalance in the system. So to affect any substantial change in the overall ratio, $\frac{Z_{pm} Z_{nm}}{Z}$ should be at least 5% of Z_{pm} i.e. 20 times Z_{pm}. But Z_{pm} it self is small (typical values in the example of 50 HP motor), so if we chose **Z** as purely resistive it causes large power drain. Inductors and capacitors for this rating become very costly and hence we resort to a N.S. sensing network without external impedance (Z).

Now from the test of 5 HP m/c we have the following data

Applied voltage = 400 Volts.

when one phase 'a' is opened

V_{ab} = 356 Volts.

V_{ac} = 380 Volts. , Input no load current = 5.5Amp.

To calculate positive and negative seq. impedances of the motor. The following data is obtained.

No load test on motor

Input volts = 300 V

Input current = 2 Amp. Speed = 1470 rpm.

Input power = 100 Watts (hundred Watts)

Slip = .02

Blocked rator test

Input voltage = 40 V

Input current = 7.5 Amp.

Input power = 180 Watt.

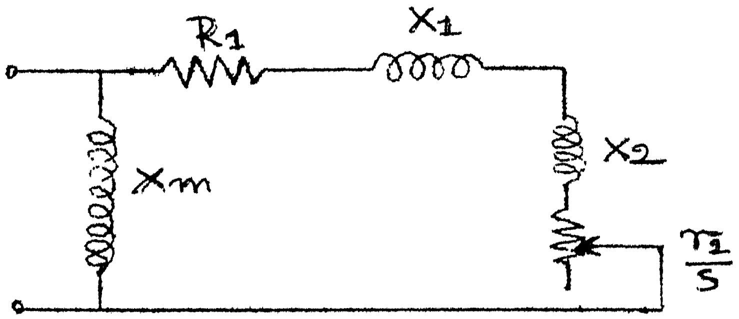


Fig 11.

Assuming the whole current goes to roter in fig 11.

$$(R_1 + r_2) = \frac{180}{(7.5)^2} = 3.25\,\Omega$$

$$R_1 \simeq 1.6\,\Omega \quad , \quad r_2 \simeq 1.6\,\Omega$$

$$V = j\,I\,(X_1 + X_2) + I\,(R_1 + r)$$
$$= j\,7.5\,(X_1 + X_2) + 12$$

Therefore, $|V|^2 = 12^2 + [7.5\,(X_1 + X_2)]^2$

$$|40|^2 = (12)^2 + [7.5\,(X_1 + X_2)]^2$$

Therefore, $|X_1 + X_2| = 5\,\Omega$

From No load test

$$\frac{r_2}{s} \simeq \frac{1.6}{.02} = 80\Omega$$

Therefore, $I^2 R = 100$

$$I^2 = \frac{100}{80} = 1.125$$

$$I = 1.06 \text{ Amp.}$$

So, $2 = \sqrt{(1.06)^2 + I_m^2}^{1/2}$

or $2^2 - 1.125 = I_m^2$

$$I_m = 1.7 \text{ Amp.}$$

hence, $X_m = \frac{300}{1.7} = 176\,\Omega$

$$Z_{pm} \simeq X_m'' \frac{r_2}{s}$$

$$= \frac{j\,176 \times 80}{80 + j\,176} = \frac{j\,176 \times 80}{37376}(80 - j\,176)$$

$$= .3767\,(176 + j\,80)$$

$$Z_{pm} = (66 + j\,30)$$

$$Z_{nm} - 3.2 + j\,5.0$$

Now we substitute the values of Z_{pm} and Z_{nm} in the following equations to calculate open phase voltages.

$$V_{ca} = \frac{\alpha^2 Z_{pm} + \alpha Z_{nm}}{Z_{pm} + Z_{nm}}$$

$$= \frac{\left(\frac{-1}{2} - j\frac{\sqrt{3}}{2}\right)(66 + j\,30) + \left(\frac{-1}{2} + j\,\frac{\sqrt{3}}{2}\right)(3.2 + j\,5.0)}{69.2 + j\,35} V_{bc}$$

$$= \frac{\frac{-1}{2}(69.2 + j\,35) - j\frac{\sqrt{3}}{2}[\,63 + j\,25]}{[69.2 + j\,35]} V_{bc}$$

$$= \frac{-35 - 17.5\,j + 21.6 - 53.7\,j}{69.2 + j\,35} V_{bc}$$

$$= \frac{-13 - 71\,j}{69.2 + j\,35} V_{bc}$$

$$|V_{ca}| = \frac{72.18}{77.54} \quad V_{bc}$$

$$= .93 \times 400 = 372 \text{ Volts.}$$

$$V_{ab} = \frac{(\frac{-1}{2} + j\frac{\sqrt{3}}{2})(66 + j\,30) + (\frac{-1}{2} - j\frac{\sqrt{3}}{2})(3.2 + 5\,j)}{69.2 + j\,35} V_{bc}$$

$$= \frac{\frac{-1}{2}(69.2 + j\,35) + j\frac{\sqrt{3}}{2}(63 + j\,25)}{69.2 + j\,35} V_{bc}$$

$$= \frac{-35 - 17.5\,j - 21.6 + 53.7\,j}{69.2 + j\,35} V_{bc}$$

$$|V_{ab}| = \frac{-56.6 + 36j}{69.2 + j35} = \frac{67}{77.54} \; |V_{bc}|$$

$$= 350 \text{ Volts.}$$

The observed values for V_{ca} = 380 volts and V_{ab} = 356 volts. So the results closely match the practically observed value.

To calculate the value of N.S. voltage practically observed we proced as follows. In the triangle below after scaling by 100,

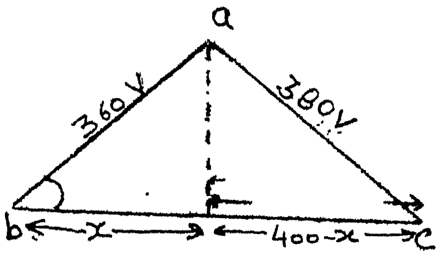


Fig 12

$$(3.6^2 - x^2)^2 = (3.8)^2 - (4-x)^2$$

$$= (3.8)^2 - 16 - x^2 + 8x$$

$$8x = (3.6)^2 - (3.8)^2 + 16$$

$$x = \frac{(3.6)^2 - (3.8)^2 + 16}{8}$$

$$= 1.815$$

so x = 181.5 Volts

$\text{Cos}\theta = \frac{1.815}{3.6} = .504166$

so, $\theta = 60^\circ$

Vector V_{cb} is shifted by 30° and V_{ab} by -30° so they overlap.

hence, N.S. voltage $= \frac{400 - 360}{\sqrt{3}} = \frac{40}{\sqrt{3}}$ volts.

which is 5% of line to line voltage.

The input current of 5 HP motor rated for 7.5 Amp. is 5.5 Amp. under such condition.

This suggests that N.S. circuit should be capable of tripping the relay when N.S. voltage is more than 5% . This physically implies that relay trips when the voltage of two phases is 380 V each and of third phase is 400 volts.

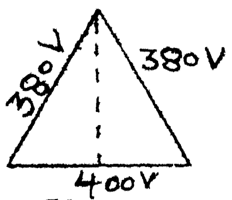


Fig 13

N.S. voltage $\simeq$ 20 volts = 5% . Fig13

CHAPTER III

Design of Sequence Filters

Since ground is generally not available with three phase systems it is better to base the designs on phase-phase quantities rather than Phase-Neutral quantities. This results in considerable cost reduction due to absence of neutral wire. Electronicaly the amplifying devices are not strained due to switching of other equipments which have neutral return path causing transients in neutral voltage. This increases the reliability of overall system and reduction in maloperation of amplifying and switching devices.

Taking phase quantities into consideration.

(1) $$V_0 = \frac{\bar{V}_{ac} + \bar{V}_{ba} + \bar{V}_{cb}}{3} = 0 \text{ (due to closed triangle)}$$

(2) $$V_1 = \frac{\bar{V}_{ac} + a^2 \bar{V}_{ba} + a \bar{V}_{cb}}{3}$$

(3) $$V_2 = \frac{\bar{V}_{ac} + a \bar{V}_{ba} + a^2 \bar{V}_{cb}}{3}$$

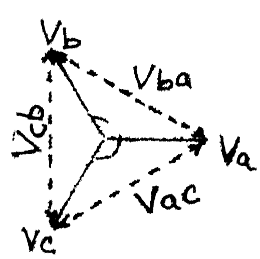


Fig 14.

$$a = e^{j} \angle 120^\circ \qquad a^2 = e^{-j} \angle 120^\circ$$

Substituting from (1) $V_{ac} = -(V_{ba} + V_{cb})$ in (2)

$$V_1 = \frac{-(V_{ba} + V_{cb}) + a^2V_{ba} + aV_{cb}}{3}$$

$$= \frac{V_{ba}(a^2 - 1) + V_{cb}(a - 1)}{3} \qquad \ldots\ldots (4)$$

Solving,

$$a^2 - 1 = \sqrt{3}\, e^j \angle -150 = \sqrt{3}\, e^j \angle 210$$

$$a - 1 = \sqrt{3}\, e^j \angle 150$$

Substituting in (4)

$$V_1 = \frac{V_{ba}\, e^j \angle 210 + V_{cb}\, e^j \angle 150}{\sqrt{3}}$$

$$= \frac{V_{ab}e^j \angle 30^\circ + V_{bc}e^j \angle -30^\circ}{\sqrt{3}} = \frac{V_{ab}e^j \angle 30^\circ - V_{cb}e^j \angle -30^\circ}{\sqrt{3}}$$

Similarly

$$V_2 = \frac{-(V_{ba} + V_{cb}) + aV_{ba} + a^2V_{cb}}{3}$$

$$= \frac{(a - 1)\bar{V}_{ba} + (a^2 - 1)\bar{V}_{cb}}{3}$$

Solving,

$$(a - 1) = \sqrt{3}\, e^j \angle 150$$

$$(a^2 - 1) = \sqrt{3}\, e^j \angle 210$$

So,

$$V_2 = \frac{V_{ba}\, e^j \angle 150^\circ + e^j \angle 210^\circ\, V_{cb}}{\sqrt{3}}$$

$$= \frac{V_{ab}e^j \angle -30^\circ + V_{bc}e^j \angle 30^\circ}{3}$$

$$= \frac{V_{ab}e^j \angle -30 - V_{cb}e^j \angle 30^\circ}{\sqrt{3}}$$

Hence steady state sequence quantities are :-

$$\boxed{\begin{aligned} V_0 &= 0 \\ V_1 &= \frac{V_{ab}e^{j\angle 30^\circ} - V_{cb}e^{j\angle -30^\circ}}{\sqrt{3}} \qquad \ldots.(5) \\ V_2 &= \frac{V_{ab}e^{j\angle -30^\circ} - V_{cb}e^{j\angle 30^\circ}}{\sqrt{3}} \qquad \ldots.(6) \end{aligned}}$$

w.r.t. phase V_{ac}

The advantage of equs. (5) and (6) is that they are in the difference form of vectors V_{ab} and V_{cb} shifted in phase by 30°. Hence these quantities can be monitered by any measuring device connected across these two vectors. This avoids need of a summer. But the disadvantage is that in equs.(5) and (6) the same vectors i.e. V_{ab} and V_{cb} are shifted in phase opposition so to derive positive Seq. and Negative Seq quantities each vector requirs two networks to shift it by $+ 30^\circ$ and $- 30^\circ$. Hence four phase shifting networks are required.

Since we are interested in the magnitudes of + ve and - ve sequence quantities and not their relative phase we can chose different reference phasors for these two quantitives.

Say we chose V_{ac} as reference for the positive sequence quantities so,

$$V_1 = \frac{V_{ab}e^{j\angle 30^\circ} - V_{cb}e^{j\angle -30^\circ}}{\sqrt{3}} \qquad \ldots. (7)$$

Now we chose V_{ba} as reference for - ve sequence quantity so,

$$V_2 = \frac{V_{ba} + aV_{cb} + a^2V_{ac}}{3}$$

$$= \frac{-(V_{ac} + V_{cb}) + aV_{cb} + a^2V_{ac}}{3}$$

$$= \frac{(a^2 - 1)\,V_{ac} + (a - 1)\,V_{cb}}{3}$$

$$= \frac{V_{ac}e^{j\angle 210} + V_{cb}e^{j\angle 150}}{\sqrt{3}}$$

$$= \frac{V_{ca}e^{j\angle 30} + V_{bc}e^{j\angle -30}}{\sqrt{3}}$$

$$= \frac{V_{bc}e^{j\angle -30} - V_{ac}e^{j\angle 30}}{\sqrt{3}}$$

$$= \frac{V_{ca}c^{j\angle 30} - V_{cb}e^{j\angle -30}}{\sqrt{3}} \qquad \ldots\ldots \quad (8)$$

From equs. (7) and (8) it is now seen that $V_{cb}e^{j\angle -30}$ is common. Only one extra phase shifting network is required to shift the phase of V_{ca} by $+ 30^{\circ}$. For convenience we take

$$\boxed{V_2 = \frac{V_{ab}e^{j\angle -30} - V_{cb}e^{j\angle 30}}{\sqrt{3}}}$$

from V_{ac} as reference and

$$V_1 = \frac{V_{ac}e^{j\angle 150} + V_{cb}e^{j\angle 210}}{\sqrt{3}}$$

$$= \frac{V_{ca}e^{j\angle -30^{\circ}} + V_{bc}e^{j\angle 30^{\circ}}}{\sqrt{3}}$$

$$= \boxed{\frac{V_{ca}e^{j\angle -30^{\circ}} - V_{cb}e^{j\angle 30^{\circ}}}{\sqrt{3}}}$$

with V_{ab} as reference

Synthesis of phase shifting networks :-

We have to shift the phase of each vector by + 30^{0} or - 30^{0}. We may use any scaling factor derived from the network impedances. Since the phase shift is $\angle 90^{0}$ we can use only passive synthesis

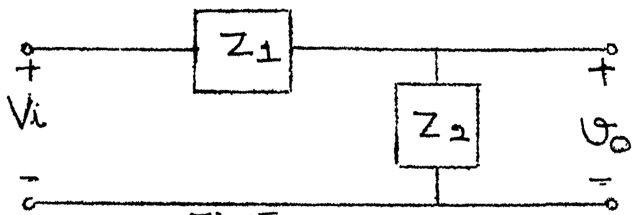


Fig 15

fig 15

$$\frac{V_o}{V_i} = \frac{Z_2\,(jw)}{Z_1 + Z_2} = \frac{|V_i|\, e^{-j\,\angle 30^{0}}}{K\,|V_i|}$$

where $V_o = \frac{V_i}{K}\, e^{-j\,\angle 30}$

K is pure real scaling number

$$\frac{Z_2\,(jw)}{Z_1\,(jw) + Z_2\,(jw)} = \frac{10}{K}\left(\frac{\sqrt{3}}{2} - j\,\frac{1}{2}\right)$$

Chosing $Z_1\,(jw) = R$

$$\frac{Z_2}{R + Z_2} = \frac{1}{K}\left(\frac{\sqrt{3}}{2} - j\,\frac{1}{2}\right)$$

or $Z_2\left[K - \frac{\sqrt{3}}{2} + j\,\frac{1}{2}\right] = .R\left[\frac{\sqrt{3}}{2} - j\,\frac{1}{2}\right]$

or $$Z_2 = \frac{R\left[\frac{\sqrt{3}}{2} - \frac{j}{2}\right]}{\left[K - \frac{\sqrt{3}}{2} + j\,\frac{1}{2}\right]} \qquad K > 1$$

$$Z_2 = R\,\frac{\left[K\,\frac{\sqrt{3}}{2} - 1\right]}{(K - \frac{\sqrt{3}}{2})^2 + (\frac{1}{2})^2} - \frac{\left[j\,\frac{K}{2} - 1\right] R}{(K - \frac{\sqrt{3}}{2})^2 + \frac{1}{4}}$$

Making $$\frac{K\,\frac{\sqrt{3}}{2} - 1}{(K - \frac{\sqrt{3}}{2})^2 + \frac{1}{4}} = 1$$

$$K = \frac{\frac{3\sqrt{3}}{2} \pm \sqrt{[\frac{27}{4} - 8]}}{2}$$ not possible due to complexroots

Chosing $$\frac{K\frac{\sqrt{3}}{2} - 1}{(K - \frac{\sqrt{3}}{2})^2 + \frac{1}{4}} = \frac{1}{2}$$

$$K^2 + \frac{3}{4} - \sqrt{3}K + \frac{1}{4} = \sqrt{3}K - 2$$

or $K^2 - 2\sqrt{3}K + 3 = 0$

$$K = \frac{2\sqrt{3} \pm \sqrt{12 - 12}}{2} = \sqrt{3}$$

$$Z_2\ (jw) = \frac{R}{2} - j\,R.\left[\frac{\frac{\sqrt{3}}{2}}{1}\right]$$

$$= \frac{R}{2} - j\frac{\sqrt{3}}{2}R = \frac{R}{2} + \frac{1}{j \times \frac{2}{\sqrt{3}R}}$$

This is a series combination of resistance and capacitance

$r + \frac{1}{jwc}$

$wC = \frac{2}{\sqrt{3}R}$

$$C = \frac{2}{\sqrt{3} \times R \times w} = \frac{2}{2\pi \times 50 \times \sqrt{3} \times R} \quad \ldots. (9)$$

Chosing $\frac{R}{2} = 120$ K

or R = 240 K

$$C = \frac{2}{2\pi \times 50 \times \sqrt{3} \times 240 \times 10^3} = \frac{1}{2\pi \times 50 \times \sqrt{3} \times 120 \times 10^3}$$

$$= \frac{1}{2\pi \times 6 \times \sqrt{3} \times 10^6} = \frac{10^{-6}}{12 \times \sqrt{3}}\,\mu\text{Farad}$$

$= .0153\,\mu$farad. (micro-farad)

So the network becomes, as shown in fig 16

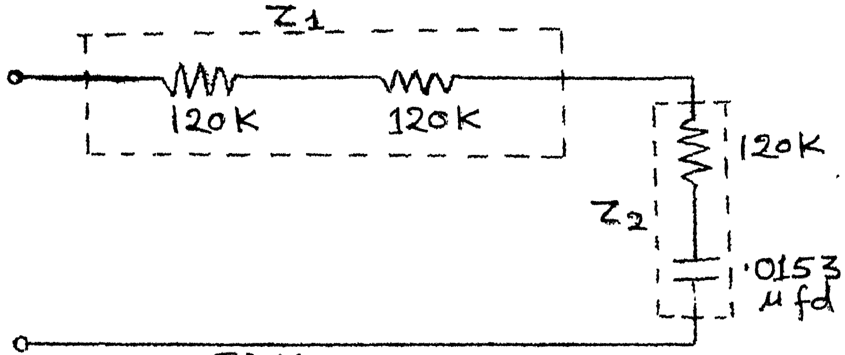


Fig 16.

Another choice would be C = .01 μfd/600V (as it is easily available) From (9)

$$R = \frac{2}{\sqrt{3} \times w \times C} = \frac{2}{\sqrt{3} \times 2\pi \times 50 \times .01 \times 10^{-6}}$$

$$= \frac{10^3}{\sqrt{3} \times \pi \times .5 \times 10^{-3}} = 367 \text{ K}$$

$$\frac{R}{2} \simeq 183 \text{ K}$$

and $wRC = \frac{2}{\sqrt{3}}$

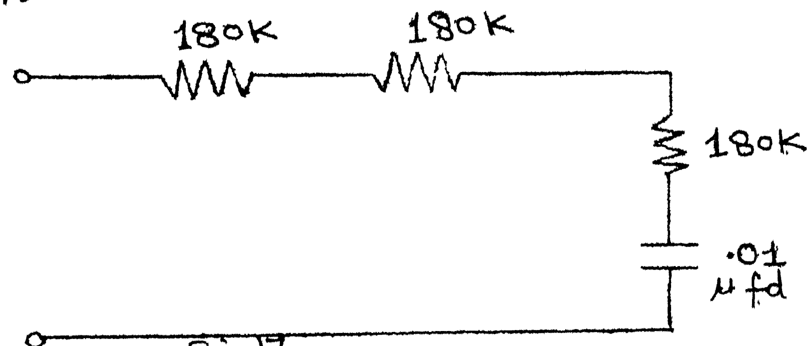


Fig 17.

Now we have to synthesize a network for rotating the input vector by $+30^\circ$.

$$T(jw) = \frac{V_o}{V_i} = \frac{V_i e^{+j \angle 30}}{KV_i} = \frac{e^{j \angle 30}}{K}$$, where $V_o = \frac{V_i e^{j \angle 30^\circ}}{K}$ and K is a scalar.

$$\frac{Z_2}{Z_1 + Z_2} = \frac{e^{j \angle 30}}{K}$$

Now chosing $Z_2 = R$

$$\frac{R}{Z_1+R} = \frac{e^{j\angle 30}}{K}$$

or $KR = (Z_1 + R) \left[\frac{\sqrt{3}}{2} + j\frac{1}{2} \right]$

Therefore, $Z_1 \left[\frac{\sqrt{3}}{2} + j\frac{1}{2} \right] = R \left[K - \frac{\sqrt{3}}{2} - j\frac{1}{2} \right]$

$$Z_1 = \frac{R \left[K - \frac{\sqrt{3}}{2} - \frac{j}{2}\right]}{\left[\frac{\sqrt{3}}{2} + j\frac{1}{2} \right]}$$

$$= R \frac{\left[K - \frac{\sqrt{3}}{2} - \frac{j}{2}\right] \left[\frac{\sqrt{3}}{2} \quad \frac{j}{2}\right]}{\left[\frac{3}{4} + \frac{1}{4} \right]}$$

$$= R \left[\frac{\sqrt{3}}{2} K - \frac{3}{4} - j\frac{\sqrt{3}}{4} - j\frac{K}{2} + \frac{\sqrt{3}}{4} j - \frac{1}{4}\right]$$

$$= R \left[\frac{\sqrt{3}}{2} K - 1 - j\frac{K}{2} \right]$$

$$= R \left[\frac{\sqrt{3}}{2} K - 1 \right] - j\frac{K}{2} R$$

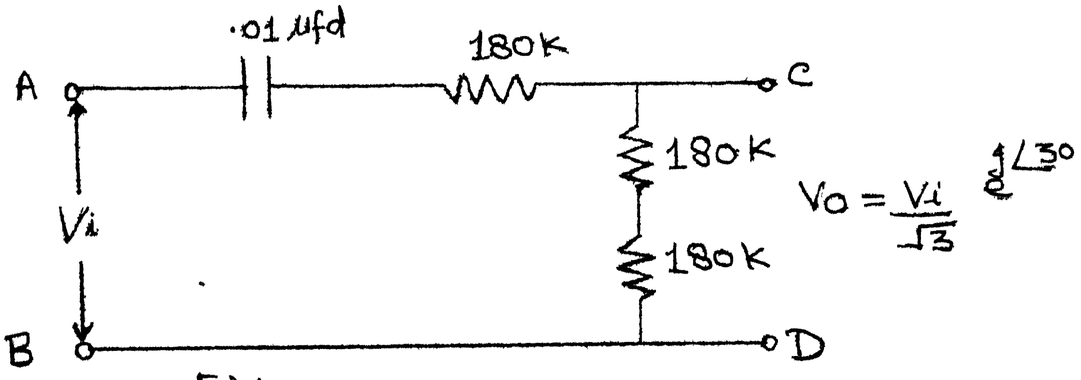
Chosing $\frac{\sqrt{3}}{2} K - 1 = \frac{1}{2}$

$K = \frac{3}{2} \times \frac{2}{\sqrt{3}} = \sqrt{3}$ we get the same amplification factor $\frac{1}{\sqrt{3}}$ as in phase lag network

$Z_1 = \frac{R}{2} - j \frac{\sqrt{3}}{2} R$

This equation is same as derived for Z_2 in phase lag network.

chosing $\frac{R}{2} = 183$ K.



$$V_o = \frac{V_i}{\sqrt{3}} e^{j} \underline{/30^{o}}$$

Hence the same network acts as a phase lead network between B and C and phase larg network between A and B with same attenation of $\sqrt{3}$. in fig 18

Practical circuit taking into account 10% tolerance is capacitor values would be as shown below :

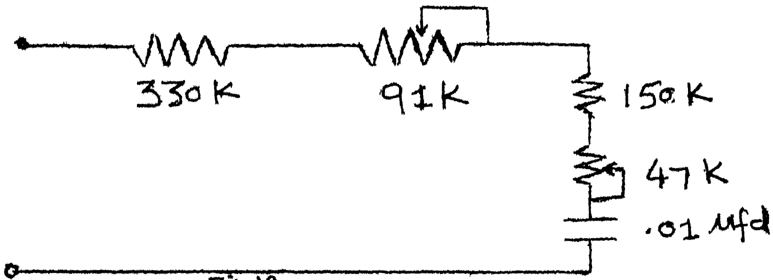


Fig 19.

Transfer function of the network

$$T(s) = \frac{Z_2}{Z_1 + Z_2} = \frac{R + \frac{1}{SC}}{2R + R\frac{1}{SC}}$$

$$= \frac{SC\ R + 1}{3\ SC\ R + 1} = \frac{1}{3}\ \frac{S + \frac{1}{RC}}{S + \frac{1}{3RC}}$$

Step responce of the above circuit is now derived.

$$V_o(s) = \frac{1}{S}\ \frac{(SC\ R + 1)}{3SC\ R + 1} = \frac{1}{3S}\ \frac{(S + \frac{1}{RC})}{(S + \frac{1}{3RC})}$$

$$= \frac{A}{S} + \frac{B}{S + \frac{1}{3RC}}$$

$$= \frac{1}{S} - \frac{2RC}{1 + 3\ SC}$$

$$\text{Therefore, } v_o(t) = u_o(t) - \frac{2}{3} e^{-t/3RC}$$

$$= \frac{1}{3} u_o(t) + \frac{2}{3}\left(1 - e^{-t/3RC}\right) u_o$$

$$\tau = 3RC = 3 \times 180 \times 10^3 \times .01 \times 10^{-6}$$

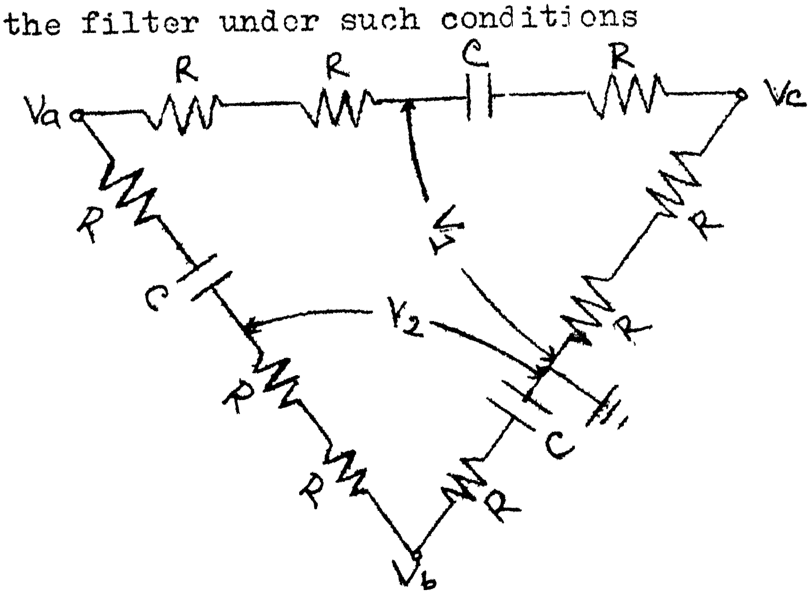
$$= 18 \times .3 \times 10^{-3}$$

$$= 5.4 \text{ m second.}$$

Assuming any sudden change in input voltage the circuit settles down in about 3 τ i.e. about 16 m second, hence the response time is well within one cycle and acceptable from practical considerations.

Frequency Sensitivity Analysis of Negative Sequence Filter :

Since the phase shifting is achieved by lumped elements, the exact phase shift of + 30° and - 30° is achieved at 50 cycles only and any variation in frequency causes both amplitude and phase error in the output of Negative Sequence filter. Since generally a variation of ± 4% is observed in frequency of mains supply it is necessary to incorporate proper compensation to avoid error in the output of the filter under such conditions



in fig 20.

$$V_2 = \frac{2R}{3R + \frac{1}{SC}} V_{ab}(s) - \frac{R + \frac{1}{SC}}{3\ R + \frac{1}{SC}} V_{cb}(s)$$

$$V_2 = \frac{2\ SCR}{3\ SCR + 1} V_{ab}(s) - \frac{SCR + 1}{3\ SCR + 1} V_{cb}(s)$$

$$\frac{\delta V_2}{\delta S} = \frac{2\ RC\ (\ 1 + 3\ SCR\ ) - 3\ RC\ x\ 2\ SCR}{(\ 1 + 3\ SCR\ )^2} V_{ab}(s)$$

$$- \frac{RC\ (\ 1 + 3\ SCR\ ) - 3\ RC\ (\ 1 + SCR\ )}{(\ 1 + 3\ SCR\ )^2} V_{cb}(s)$$

$$= \frac{2\ RC}{(\ 1 + 3\ SCR\ )^2} V_{ab}(s) + \frac{2\ RC}{(\ 1 + 3\ SCR)^2} V_{cb}(s)$$

$$\frac{dV_2}{\delta S} = \frac{2\ RC}{(\ 1 + 3\ SCR)^2} (V_{ab}(s) + V_{cb}(s))$$

$$\left.\frac{\delta V_2}{\delta S}\right|_{s = jw} = \frac{2\ RC}{(\ 1 + 3\ jWRC)^2} [\ V_{ab}(jw) + V_{cb}(jw)\ ]$$

$$= \frac{2\ RC}{(\ 1 - 9\ W^2R^2C^2) + 6\ jWRC} [\ V_{ab}(jw) + V_{cb}(jw)]$$

Therefore, $$\left.\frac{\delta V_1}{\delta S}\right|_{s = jw} = \frac{2\ RC}{(1 - 9\ W^2R^2C^2) + j\ 6\ RCW} (\bar{V}_{ab} + \bar{V}_{ab}) \quad \ldots.\ (11)$$

From Equ.(10) the components are chosen so that $WRC = \frac{1}{\sqrt{3}}$,

substituting in (11)

$$SWRC = \frac{1}{\sqrt{3}} \text{ and } RC = \frac{1}{w\sqrt{3}}$$

$$\frac{\delta V_1}{\delta S} \quad s = j\ w \quad \frac{2\ RC\ (\bar{V}_{ab} + \bar{V}_{cb})}{(1 - 9\ x\ \frac{1}{3}) + j\frac{6}{\sqrt{3}}} = \frac{2\ (\bar{V}_{ab} + \bar{V}_{cb})}{\sqrt{3}\ W\ (-2 + j\frac{6}{\sqrt{3}})}$$

$$= \frac{(\ \bar{V}_{ab} + \bar{V}_{cb})}{\sqrt{3}\ W\ [-1 + j\ \sqrt{3}]}$$

$$\therefore \left| \frac{\delta V_1}{\delta f} \right| = \frac{1}{\sqrt{3}\ f \times 2} \left| V_{ab} + V_{cb} \right|$$

$$= \frac{1}{\sqrt{3}\ f \times 2} \times \sqrt{3} \left| V_{ab} \right|$$

$$\left| \frac{\delta V_1}{\delta f} \right|_{peak} = \frac{\sqrt{3}\ V_{ab}}{2\ f} = \frac{\sqrt{3} \times 220 \times \sqrt{3} \times \sqrt{2}}{100}$$

$$= 9.25 \text{ Volts/Hz.}$$

Since we want to trip the motor on 5% imbalance which is $400 \times \frac{\sqrt{3} \times 5}{100} = 20.0 \times 1.73 = 34.5$ voltase, peak a four percent variation in frequency i.e. 2 Hz will give approximately $2 \times 9.25 = 18.5$ voltase, signal.

This causes appreciable error under balanced condition with 4% frequency deviation although the motor may be capable of running at these frequencies without exceeding magnetizing and load current limitations.

To avoid this problem a frequency compensation circuit is designed which adds voltage in opposite polarity proportianal to frequency and cancels out the frequency effect from this circuit.

Phase Error Analysis of the Negative Sequence Network :

Considering only one branch of N.S. network in fig 21

$$v_o = \frac{2\ RCS}{1 + 3\ RCS} V_i$$

Therefore,

$$\theta = \pi/2 - \tan^{-1} 3\ WRC$$

$$\frac{d\theta}{dW} = \frac{1}{1 + 9\ W^2R^2C^2} \times 3\ RC$$

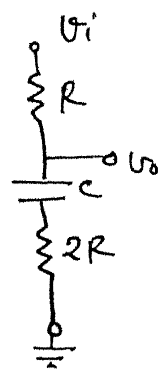


Fig 21.

$$= - \frac{3 \ WRC}{W \left[1 + 9 \ W^2R^2C^2 \right]}$$

$$\left| \frac{d\theta}{df} \right| = - \frac{1}{f} \times \frac{3 \times \frac{1}{\sqrt{3}}}{1 + 9 \times \frac{1}{3}} = - \frac{1}{f} \quad \frac{\sqrt{3}}{4}$$

$$\left| \frac{d\theta}{df} \right| = \frac{1.73}{50 \times 4} \ \text{rad/Hz}$$

Therefore,

$$\therefore \Delta\theta = 2 \times \frac{1.73}{50 \times 4} \times \frac{180}{\pi} \ \text{deg}$$

$$\frac{360}{200} \times \frac{1.73}{3} \ \text{degrees for 2 cycle variation}$$

$$= \frac{1.8 \times 1.73}{3} = 1.02 \ \text{degr}$$

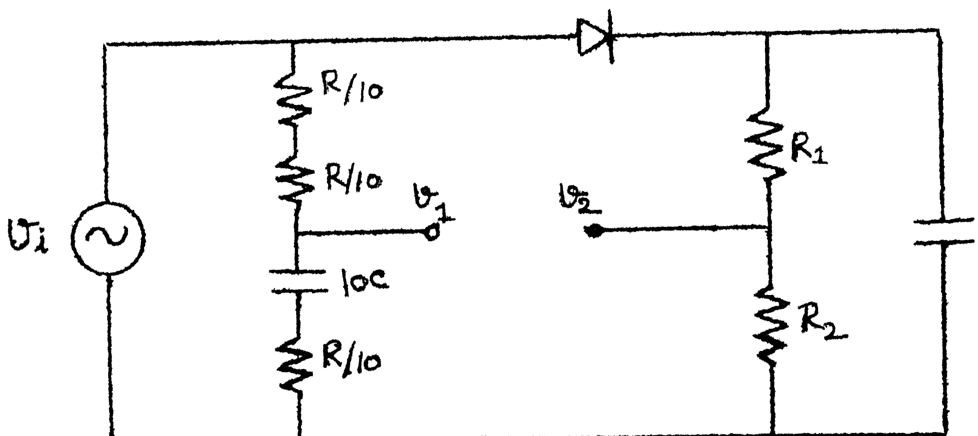
So the error in phase is 1.02 degrees i.e. 3% and we may still treat the network as true negative sequence filter within 48 - 52 Hz input.

Frequency Compensation Circuit :

We need a compensation circuit which gives us frequency sensitivity according to equation (11)

$$\text{i.e.} \left| \frac{\delta V}{\delta S} \right|_{s = jw} = \frac{2 \ RC}{(1 - 9 \ W^2R^2C^2) + j \ 6WRC} \ V_i$$

For this parpose we take a network which has the same configuration as that for determing sequence quantities but scaled by a factor say 10 as shown in figure 22 below.



Suppose we consider a input voltage ϑ_i in figure 1 then

$$\vartheta_1 = \vartheta_i \times \frac{SCR + 1}{3\ SCR + 1}$$

$$\frac{\delta V_1}{\delta S} = \frac{RC\ (1 + 3\ SCR) - 3\ RC\ (1 + SCR)}{(1 + 3\ SCR)^2}$$

$$\left|\frac{\delta V_1}{\delta S}\right|_{S=j\omega} = \frac{-2\ RC}{\left|(1 - 9\ W^2R^2C^2) + j\ 6\ WRC\right|} \left|\vartheta_i\right| \quad \ldots\ldots \quad (12)$$

So the amplitude sensitivity is same as required above.

For determining the magnitude of ϑ_1 we rectify it and do peak detection. At 50 Hz WRC $= \frac{1}{\sqrt{3}}$ and $\vartheta_1 = \frac{1\, e^{-j\angle 30}}{\sqrt{3}}$

$$\left|\vartheta_1\right| = \text{d.c.}\ \frac{\left|\vartheta_i\right|}{\sqrt{3}}\ \text{d.c.}$$

Now the ϑ_i is peak detected and the resistiros R_1 and R_2 are chosen so that

$$\frac{R_2}{R_1 + R_2} = \frac{1}{\sqrt{3}} \quad \text{so} \quad \vartheta_2 = \frac{\vartheta_i}{\sqrt{3}}$$

and $$\vartheta_1 - \vartheta_2 = \vartheta_i \left|\frac{1 + SCR}{1 + 3\ SCR}\right| - \frac{\vartheta_i}{\sqrt{3}} \quad \ldots\ldots \quad (13)$$

Difference $\left|\vartheta_1 - \vartheta_2\right|$ is taken and added in phase opposition with the negative sequence voltage derived earlier after N.S. voltage has been rectified. The output of the adder gives the total N.S. voltage. as shown in fig 23.

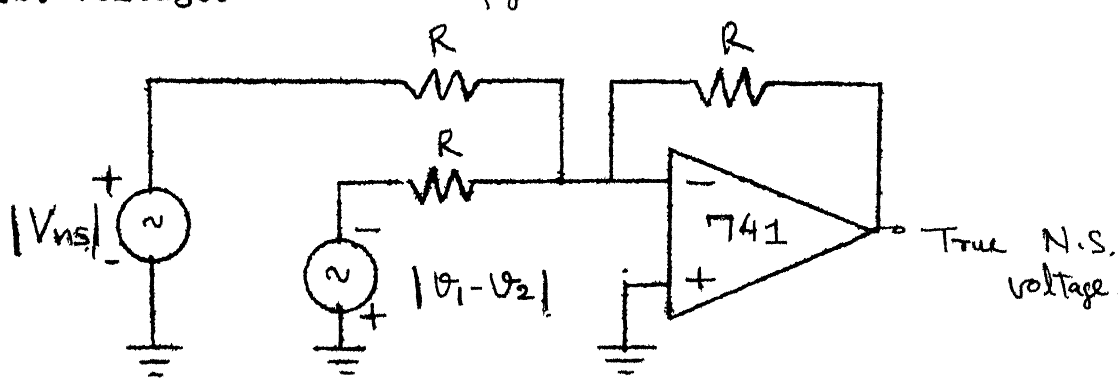


Fig 23.

The first advantage of above scheme is that first term in equ.(13) is frequency dependent having the same sensitivity in amplitude as required for N.S. voltage while the second term has a constant coefficient so overall sensitivity is same as that of first term. Hence at 50 Hz the coefficients match and the compensation voltage is zero.

$$\vartheta_1 - \vartheta_2 = \vartheta_i \; T(s)_{50Hz} - \frac{\vartheta_i}{\sqrt{3}}$$

$$\vartheta_i - \vartheta_2 = \frac{\vartheta_i}{\sqrt{3}} - \frac{\vartheta_i}{\sqrt{3}} = 0$$ irrespective of input voltage amplitude.

Second advantage of this scheme is that if the three phase voltages are balanced and only the frequency is deviated then also we get true compensation and almost zero net N.S. voltage irrespective of input voltage, because

$$|V_{NS}|_{\text{true}} = |V_{NS}|_{\text{network}} - |V|_{\text{compensation}}$$

$$\frac{\delta |V_{NS}|_{\text{true}}}{\delta W} = \frac{\delta |V_{NS}|_{\text{network}}}{\delta W} - \frac{\delta |V|_{\text{compensation}}}{\delta \omega}$$

$$= 0$$

$\therefore V_{NS} = \text{Constant} = 0$ (Under balanced Condition)

For this purpose V_i should be chosen as the positive sequence voltage. But since positive sequence voltage is derived from mains the circuit will become complicated and would require further amplifiers. An easier way out is to use one of the phase to phase voltage as input which is also supplying power to whole unit through transferer. This would save the need of further rectification and filteration as rectifiers is already there to derive power supplies. This would introduce error

in compensation to the extent that if this phase becomes unbalanced it will not provide true compensation against frequency variation. But since maximum permissible unbalance is 5% taking one of the phase voltage as V_i would not introduce much error in the compensation voltage.

This circuit is simple cheap and accurate compared to any other circuit which would have been designed on reference frequency or reference voltage internally generated as they would have required very stable voltage and frequency source and also its synchronization with the input voltage.

This same circuit has been used to give trip command to the relay on overfrequency operating condition with an accuracy better than 5% i.e. $2 \times \frac{5}{100} = .1$ Cycle.

The actual implementation of the scheme is shown in chapter 5 ahead.

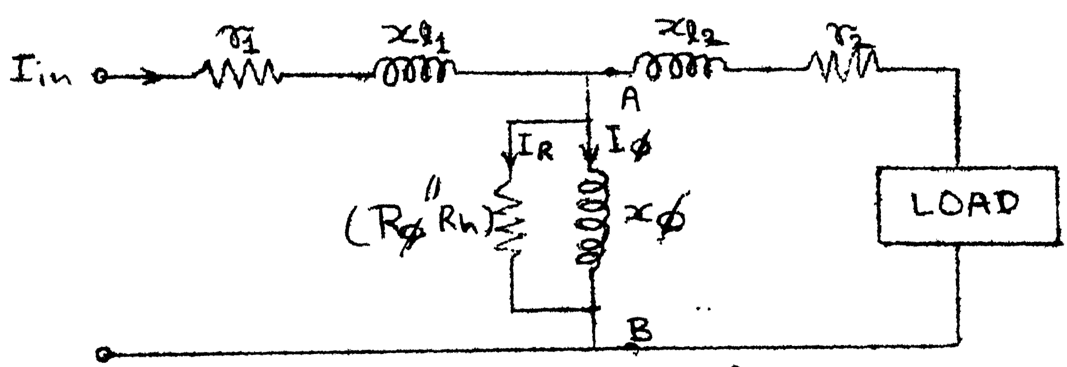
Design of Current Coils :

Till now the emphasis was placed on voltage sensing and calibrating voltage quantities using lab oscilator design. Voltage detection provides adequate protection against overvoltage, undervoltage and single phasing etc. but it fails to provide any protection against overload, low acceleration and earth fault conditions. This necessiates the need for current sesing and processing to avoid damage to motors. Further for big motors having large inertia loads, it has been derived from analysis as shown in chapter II that the voltage induced in the phase that opens out is comparable to the line voltage when motor is running near synchronous speed, under this condition of single phasing purely voltage sensing fails. If purely current sensing is utilized for single phasing or unbalance protection, this suffers from the disadvantage that current transients that might occur during loading, unloading starting etc. may cause undesirable tripping. This is especially true of motors where rapid reversal is taking place or oscilating load is connected. Negative and positive sequence filters under such conditions will generate their own transients if current sequenses are derived and will have to be deisgned for a wide variation of input quantities, which will cause error in the control

variables which are governing the final operation of the relay. Here voltage sensing is advantageous because for normal three phase supply with not very large line impedance will act as almost ideal voltage source for normal loading conditions of motor and hence the effect of current transients on voltage will be very little. Thus the negative sequence filter is not subjected to very severe voltage transients and fluctuations due to the loading of machine, hence accuracy of operations can be maintained. Therefore hybrid (voltage cum current) protection scheme is desirable from the point of view of maintaining good overall accuracy.

The basic element for current sensing, the current transformer is required for sensing zero sequence and normal load current. The attempt is made to design a low cost high accuracy current transformer which can be used as a general current sensing element with any equipment.

A short review is made of the conventional CT's.



Equivalent Circuit of CT.

Fig 24.

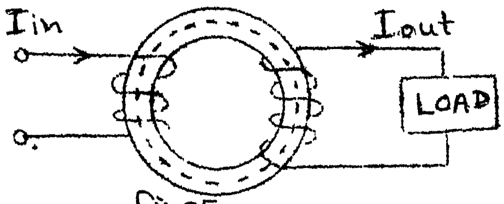


Fig 25

Normaly <u>one turn</u> primary is wound over a magnetic material with large number of secondaries feeding a load. The load normally consists of a low impedance coil of a relay or some other protective scheme. The electrical equivalent of the transformer is shown in ~~the above diagram.~~ fig 24.

<u>This CT suffers from the following inherent inaccuracies</u> :

The primary current I_{in} after entering the CT is consumed partly as hysterisis loss R_h and partly as core loss R_ϕ due to induced currents. Since secondary has finite load connected across it, there is some seconday voltage induced at output. This voltage requires that there should be some finite flux through the core to sustain the voltage. Hence magnetizing current flow through x_ϕ. Since primary is a single turn coil, this magnetizing current may be enough to cause apprecable error in the secondary current. Even if the seconday load is made zero, since secondary contains large no.of turns the leakage inductance x_{12} and r_2 may develop, due to circulation of current, quite large voltage across AB which again causes this error in the output current transformation.

Second problem associated with such construction is that at high currents (say overload) the core material may

be subjected to operate in the Nonlinear region of B.H curve and correct estimate of input current can not be made under such abnormal conditions.

Third disadvantage of such CT's arises out of their use with modern low power high gain IC's like OP Amps etc. The OP Amps say μA741 has maximum current capability of 10 m Amp. To minimize current transformation error suppose we connect this CT in the virtual ground mode as shown in the following figure 26.

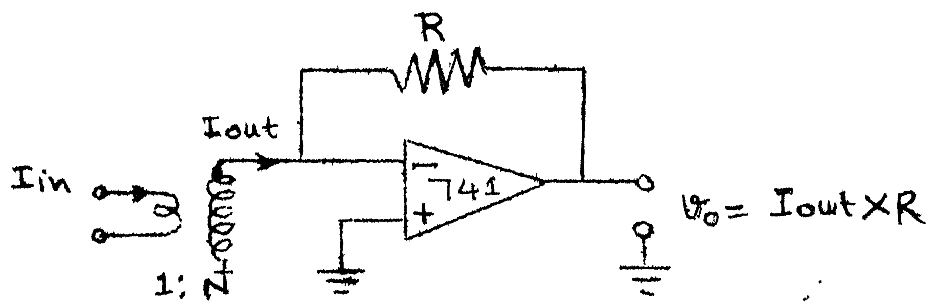


Fig 26

Now neglecting magnetizing and hysteresis loss error,

$$I_{out} = \frac{I_{in}}{N}$$

for a 10 HP motor the input current for each phase $\simeq$ 10 Amp.

Since 741 can sink or source maximum 10 m Amp current so $I\ out_{max}$ = 10 m Amp

so $\frac{10\ Amp}{N}$ = 10 m Amp

or, $N = \frac{10}{10 \times 10^{-3}} = 1 \times 10^{3}$ = One thousand turns.

For linear operation of Opamp at 100% overload i.e. about 20 Amp Two thousand turns are required.

Thus a large no.of secondary turns are required to transform the current to OP Amp compatible levels. Such a large no.of turns will cause large current transformation ratio error in the CT. Although this design may be acceptible for low input current levels.

Thus these CT's can be utilized with advantage where large secondary current (Power) is required to operate some relay etc. directly without much processing.

An alternative method is adopted to construct the CT which eliminates the above mentioned disadvantages and is cheaper than the conventional design.

<u>The step by step development procedure for this new construction is as follows</u> :

In fig 24. Suppose we open the secondary i.e. Load is disconnected, then there is no secondary current. Now the primary current flows through x_{l1}, r_1 and x_ϕ and $(R_\phi + R_h)$.

R_1 is the resistance of that portion of primary coil which is linked through magnetic circuit to secondary. This R_1 we can treat as a part of the cable through which current is flowing. The leakage reactance x_{l1} is the air gap flux (In air) which does not link secondary and has closed rings formed in air, so Xl_1 can also be treated as lead inductance of the cable carrying this current. Naturally one would argue that open secondary means high impedance to primary circuit because of the presence of the core.

But we remove the magnetic core and use some nonmagnetic material or air.

The magnetizing impedance $x_{\emptyset}$ reduces drastically now but at the same time the coupling coefficient K between primay and secondary reduces drastically.

Now the equvalent circuit becomes as shown below. in fig 27.

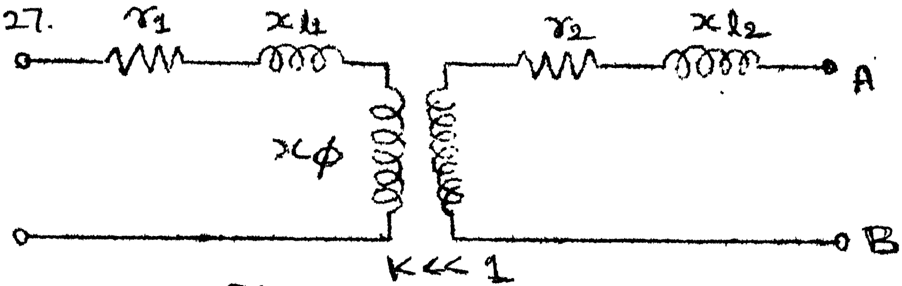
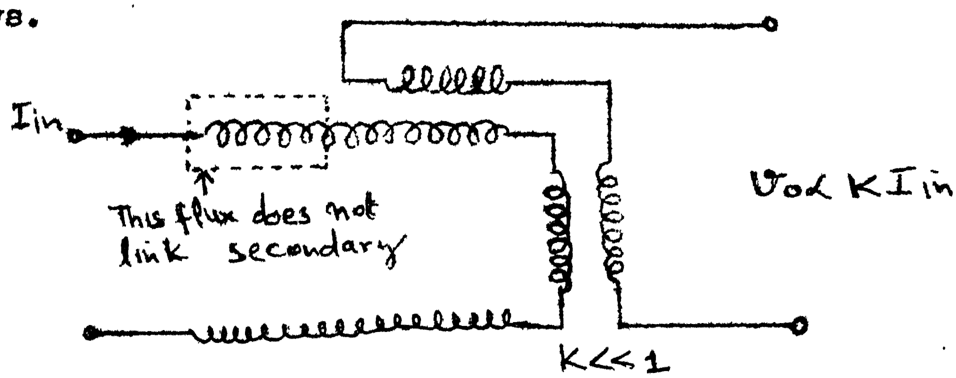
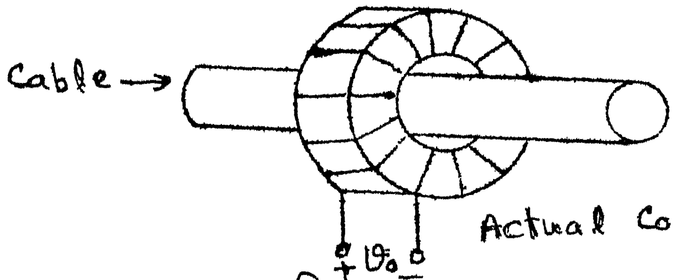


Fig 27.

Since now the core is also air so x_{11} and $x_{\emptyset}$ are of the same order of magnitude. Since the core material is air there is no loss in primary so $R_{\emptyset} + R_h = \infty$. Since there is no secondary load, so r_2 and r_{12} lose meaning and hence no magnetizing current is required to sustain secondary voltage due to load. The voltage across AB is the open circuit voltage induced due to primary-secondary coupling. x_{11} and $x_{\emptyset}$ lose their identify as leakage and magnetizing impedance because the core material is nonmagnetic. We can only say that which portion of the flux of the cable is utilized in generating the secondary voltage. Thus the equivalent circuit can be represented as follows.



Equivalent ckt of air core current



Actual Construction of CVT.

Fig 29.

The leakage flux of the current carrying cable links certain number of turns of a coil and a voltage proportional to input current is generated in this coil. The advantage here is that we can increase number of turns in secondary to quite a large extent without causing any error due to r_2 and Xl_2. Secondly there is no magnetizing current required for developing secondary voltage. Since air core in used saturation does not occur in a very large dynamic range and linearity is maintained.

No extra impedance is introduced in the primary current path.

Even for very large current transformation the cost of CT does not increase because of absence of magnetic core. The same CT can be used from 1 HP to 25 HP motor without any modification.

The advantage is now taken of high gain of µA 741 by connecting the CT in the high input impedance mode and using it as amplifier of gain 100. as shown in fig 30.

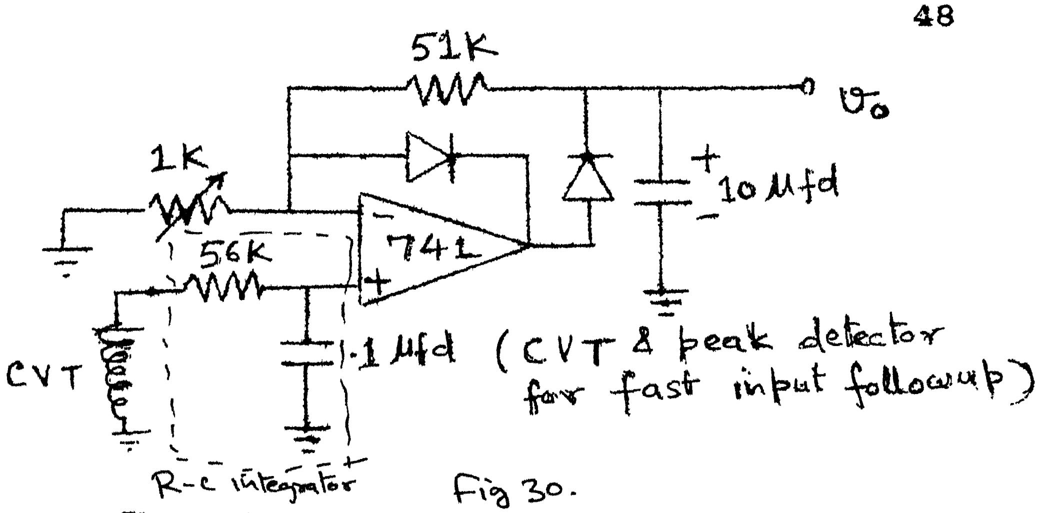
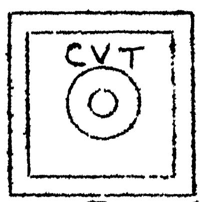


Fig 30.

The output can be normalized to any desired level for all CT's by varying 1.0 K pot. The disadvantage of this type of construction is the interference from external magnetic signal. But cast iron seperators can be provided as shown in the following figure to avoid interaction of external flux.



Cast Iron Shield

CHAPTER- 5

Actual Hardware Realization Of The Scheme :

The main transformer for feeding power to the OP Amps. is a 400Vrms/ 25 $_{peak}$Volts$_{peak}$ T/F. The DC operating voltages for OP Amps. are ± 15V so a zener regulated power supply is used. A center tap transformer is designed with secondary ± 25 V/1 Amp.

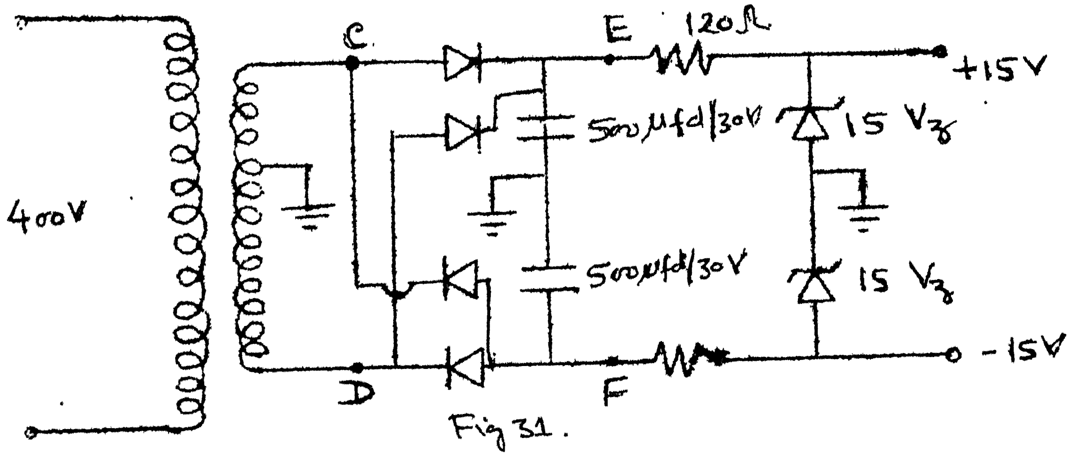


Fig 31.

The negative sequence filter is connected in the following manner shown in fig 20. One of the terminals across which V_2 (N. S. Voltage) is measured is grounded to Transformer centre tap.

The setting for 5% unbalance turns out to be $\frac{400}{\sqrt{3}} \times \sqrt{2} \times \frac{5}{100}$ Volts peak $= 16.5$ Volts peak.

Beyond this the relay should trip. Now for frequency compensation as discussed in chapter 3 the N.S. network is scaled down by a factor of 10. So $R_C = \frac{R}{10}$ and $C_C = C \times 10$.

Where R_C and C_C refer to compensation network R and C to original N.S. network.

So $C_C = .1\mu$ farad. , $R_C = \frac{180}{10} = 18K$

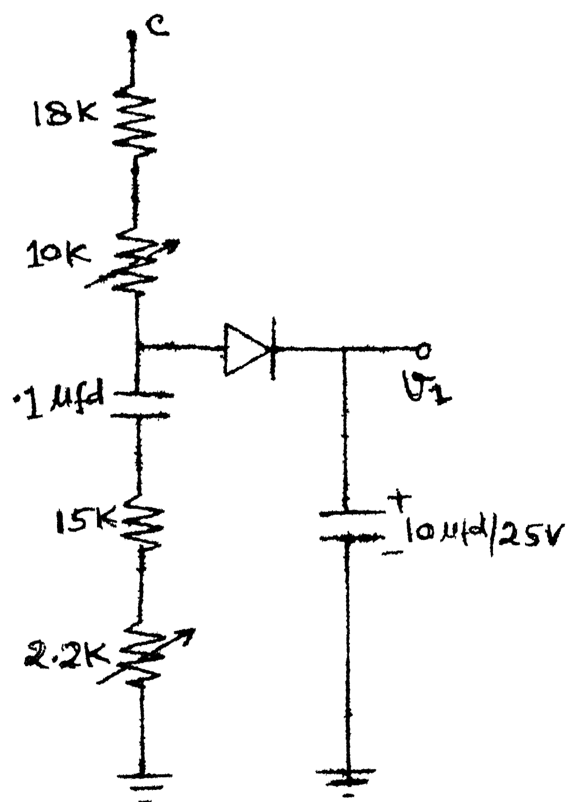


Fig 32.

The AC signal for frequency compensation is derived from point C of the power supply transformer. Fig 31 Voltage ϑ_2 is derived from negative power supply from point F. Fig 31 To get absolute value of the difference of $\vartheta_1 - \vartheta_2$ the following circuit is used.

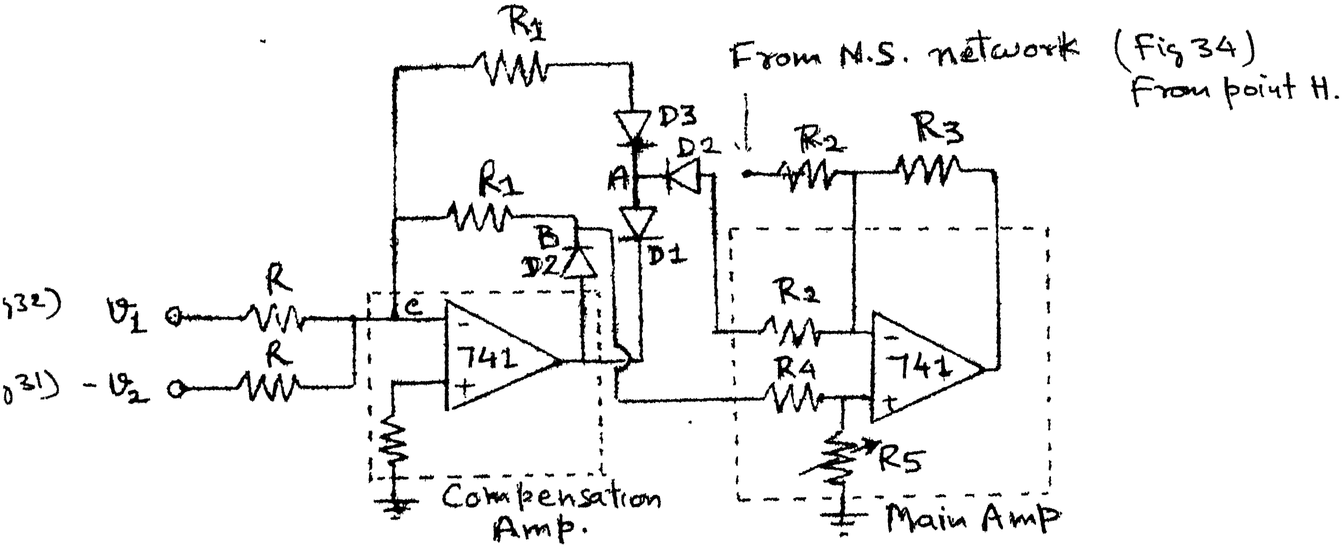


Fig 33.

ϑ_1 is the rectified voltage from compensation circuit (Fig 32) and ϑ_2 is the rectified negative voltage point F in fig 31. At 50 Hz $\vartheta_1 = -\vartheta_2$ so no current flows into the inverting input of OP Amp. and voltage at points 'A' and 'B' is zero. Main Amp. amplifies only rectified N.S. voltage. When frequency is greater than 50 Hz. $|\vartheta_1| < |\vartheta_2|$ so D_2 conducts and we get a net positive voltage at B = $+\frac{R_1}{R}(\vartheta_2 - \vartheta_1)$. The voltage at point 'A' is zero because (-ve) input is virtual ground. When frequency is $<$ 50 Hz $|\vartheta_1|$ greater than $|\vartheta_2|$ so D_1 conducts. and output voltage is $\frac{-R_1}{R}(\vartheta_1 - \vartheta_2)$ while voltage at point B is zero. A is connected to inverting input of main Amp.so the overall amplification becomes $\frac{+R_1}{R}(\vartheta_1 - \vartheta_2) \times \frac{R_3}{R_2}$ and since when A conducts voltage at B is zero so net voltage is -

$$\frac{+R_1}{R} \left|\vartheta_1 - \vartheta_2\right| \quad \frac{R_3}{R_2}, \quad (\vartheta_1 > \vartheta_2) \qquad -I$$

Since B is connected to non inverting input of main Amp. the overall amplification becomes $\frac{+R_1}{R}(\vartheta_2 - \vartheta_1)$ $\times \frac{R_5}{R_4 + R_5} \times \left(\frac{R_3}{R_{2/2}} + 1\right)$ for $(|\vartheta_2| > |\vartheta_1|)$

And when B conducts voltage at 'A' is zero so overall voltage is $\frac{R_1}{R}(\vartheta_2 - \vartheta_1) \quad \frac{R_5}{R_4 + R_5} \times \left(\frac{R_3}{R_{2/2}} + 1\right)$ -II

Hence effectively we get compensation voltage of same polarity. The advantage of this configuration over normal absolute value circuit is that for each signal we require only one input resistor and no matching is required. We are now interested in equal gain and in equns. I and II.

Equating multiplying factors in I and II

$$\frac{R_1}{R} \times \frac{R_3}{R_2} = \frac{R_1}{R}\left(1 + \frac{R_3}{R_{2/2}}\right) \times \frac{R_5}{R_4 + R_5}$$

$$\frac{R_5}{R_4 + R_5} = \frac{R_3/R_2}{1 + 2\frac{R_3}{R_2}}$$

$$\text{or } 1 + \frac{R_4}{R_5} = 2 + \frac{R_2}{R_3}$$

$$\text{or } \frac{R_5}{R_4} = \frac{R_3}{R_2 + R_3}$$

Now from equ. (11)

$$\left.\frac{\delta V_1}{\delta S}\right|_{S=j\omega} = \frac{2\ R\ C}{(1-j9\,\omega^2R^2C^2)+j6\ \ RC} \quad (\bar{V}_{ab} + \bar{V}_{cb})$$

and from equ.(12) the rate of change of compensation voltage is $\left|\frac{\delta V_1}{\delta S}\right|_{comp} = \frac{2\ R\ C}{(1-j9\,\omega^2R^2C^2)+j6\omega\,RC}\ \vartheta_i$

Now $\vartheta_i = 25\ V_{peak}$

while the negative sequence voltage is being monitored directly from mains,

$$\text{so} \quad V_{ab} + V_{cb} = \sqrt{3}\ \ V_{ab}$$

Hence peak voltage is $= \sqrt{3} \times \sqrt{3} \times \sqrt{2} \times 220$

$= 3 \times 1.4 \times 220 = 4.2 \times 220$

$= 924$ volts peak

so for true compensation these two should be brought to same level.

The ratio is $\frac{25}{924} \simeq \frac{1}{37}$

Since we need to switch off the relay at 17 V_p, to make this voltage OP Amp. compatible we reduce N.S. voltage by a factor of 4 and boost compensation voltage by 9.1 to get overall gain $4 \times 9.1 = 36.4 \simeq 37$.

The resister for ϑ_1 is chosen as 10K. To get a factor of $\sqrt{3}$ for ϑ_2 $R = 1.73 \times 10 = 17.3K$ connected directly to - V(25V). For a gain of 9.1, $R_1 = 91$ K.

Now to detect N.S. voltage the following circuit is used

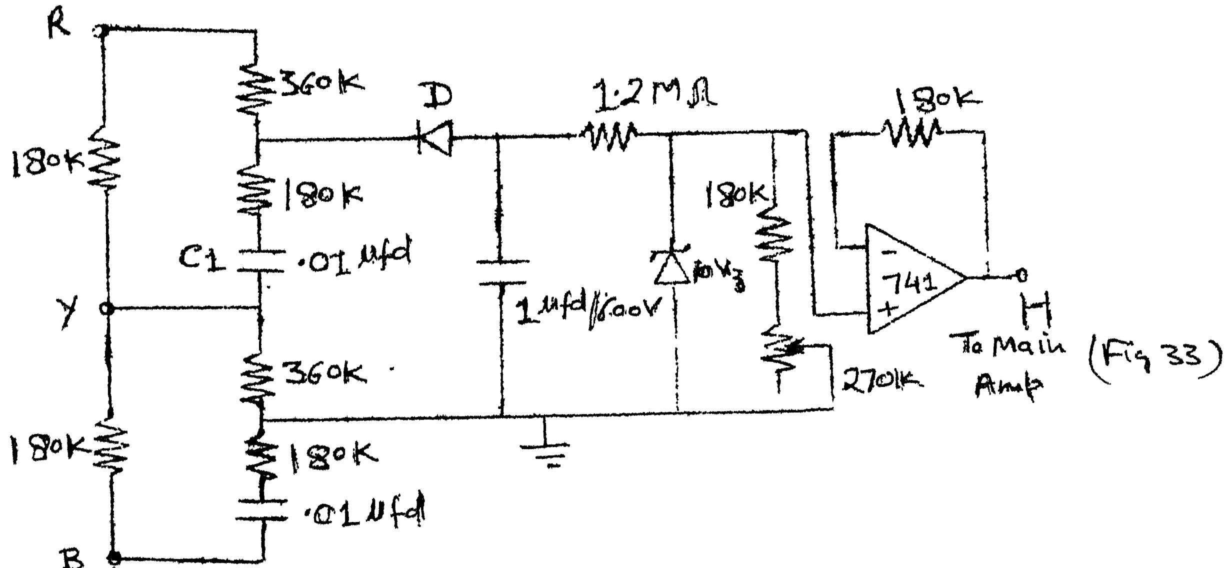


Fig 54

The advantage of above configuration is that 1 μfd/600 V capacitor gets charged to the true peak and after a scaling of 390/1590 $\simeq \frac{1}{4}$ it goes to OP Amp. which serves to act as true N.S. voltage source. 1.2 M resistor acts as a high impedance load so we get true peak detection rather than an average. The N.S. network acts as a high impedance voltage source. The approximate source resistance being 360K, which may be more depending on the type of imbalance especially when one phase opens out. Say phase 'R' opens out then current in 1 μfd capacitor flows through D and capacitor C_1= .01 μfd. So C_1 also gets charged but because of presence

of D there is no discharge path available for C_1. So to provide discharge path under such circumstances resistances R_1 = 180 K are connected across phases. These resistors do not play any role when all the three phases are connected but when any of the phase is disconnected these resistances provide discharge path to C_1's.

Suppose we derive unity gain from main Amp.

i.e. $R_3 = R_2$

so $\frac{R_5}{R_4} = \frac{1}{2}$

say R_3 = 10K, R_2 = 10K, so R_5 = 10K, R_4 = 20K.

So the final circuit takes the following form fig 35.

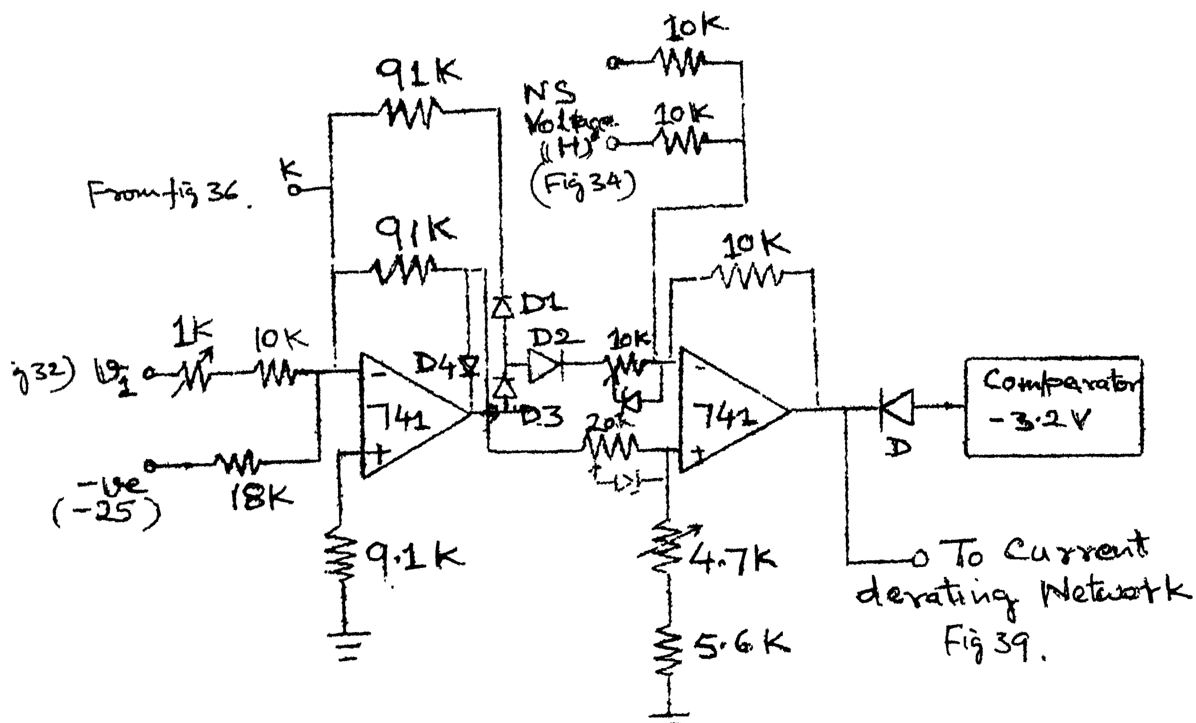


Fig 35

Diode D_1 is placed to avoid feed back to camp. Amp when D_4 conducts because when the voltage is applied to (+) input of main Amp., the voltage is fed back to comp.Amp through (-ve) input of main Amp. which is prevented by D_1.D_2 is for V_{BE} compensation of D_1. Since D_1 prevents conduction through (10K) this resistor does not play any role and acts as a open circuit hence does not contribute to the gain for any positive voltage applied at noninventing input of main Amp.

Undervoltage And Overvoltage Sensing :

Since the unit is intended to switch off either at under voltage or overvoltage, we only need to derive a signal which gives trip command to relay. This is accomplished with the aid of zeners as references. The use is made of the fact that trip signal should exist when either of the condtions is violated. In the operating region there should not exist any singnal. If this technique is adopted then we can amplify this switching signal in any existing amplifier intended for amplifying some other continuous signal. We made use of the compensation amplifier for doing this job. Another advantage of using this stage is that since it acts as a rectifier also, the polarity of the switching signal does not effect the performance.

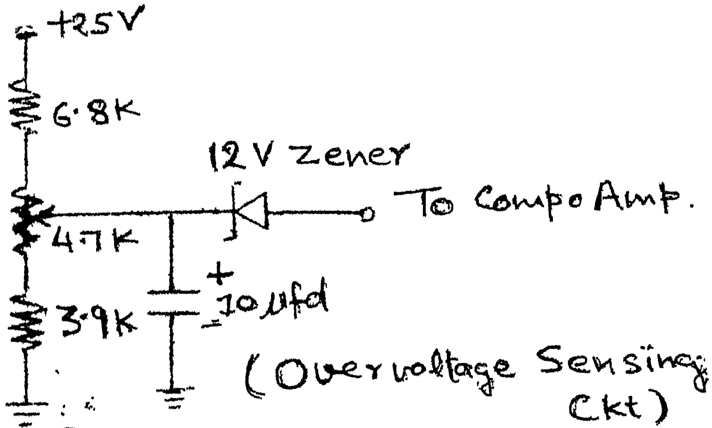


Fig36.

If the motor can bear 10% over voltage the unregulated supply will be 27.5 volts. (25+2.5). With 6.2V ~~6.12V~~ zener

$$\frac{R_2}{R_1 + R_2} = \frac{R_2}{15} \times 27.5 = 12 \text{ V}$$

$$R_2 = \frac{12 \times 15}{27.5} = 7.0 \text{ K}\Omega .$$

so when it is used with comp Amp with R_1 = 91 K the gain is $\frac{91}{7}$ = 14, hence when voltage exceds 27.5 volts by .21 volts the output is = .2 x 25 = 5.00 volts which switches the comparator. By varying the pot the adjustment for over voltage limit can be made.

For under voltage sensing the following circuit is used say at - 10% i.e. (25-2.5)V we wish to switch the ckt.

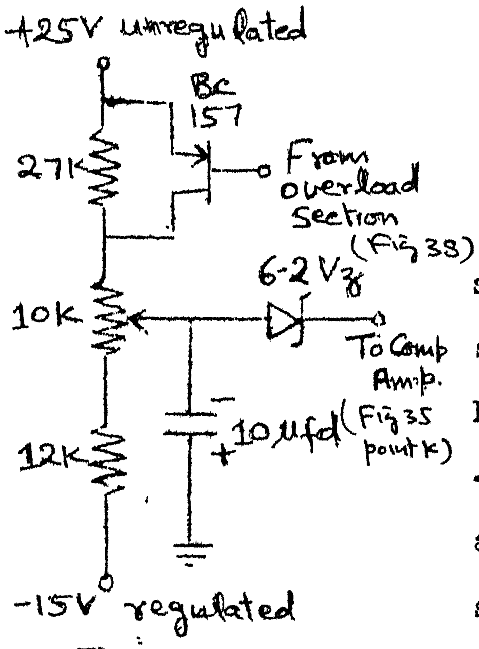


Fig 37.

$$\frac{R_2}{R_1+R_2} V + -\frac{R_1}{R_1+R_2} V^{-} =$$

6.2 volts.

$$\frac{R_2 x\ 22.5}{R_1 + R_2} - \frac{R_1 x\ 15}{R_1 + R_2} = -6.2$$

say $R_1 + R_2 = 47K$

so, $22.5\ R_2 - 15\ R_1 = -\ 290$

hence, $22.5\ (47-R_1) - 15R_1 =$
$- \ 290$

and, $37.5\ R_1 = 1290$

so, $R_1 = \frac{1290}{37.5} = 33\ K$

$R_2 = 14\ K$

With these resistors in circuit we check the voltage when input voltage is at maximum i.e. 27.5 volts.

$$27.5 \times \frac{14}{47} - \frac{15 \times 33}{47} = 9.2 - 11 = -\ 2 \text{ volts.}$$

Hence zener does not conduct as forward diode under high voltage condtions in fig 37.

The gain of this stage is $\simeq \frac{91}{14\ K''\ 33K} \simeq 9$

To avoid undervoltage tripping under starting and overload condition the PNP transistor is provided which is switched from overload section.

This completes the voltage control design of the unit. Now we proceed to current control design.

Section For Sensing Overload And Locked Rotor Condition:

The overload is sensed by a CT and amplified and rectified voltage is obtained at a normalized level of .5 volt at rated load by varying R_1 as following.

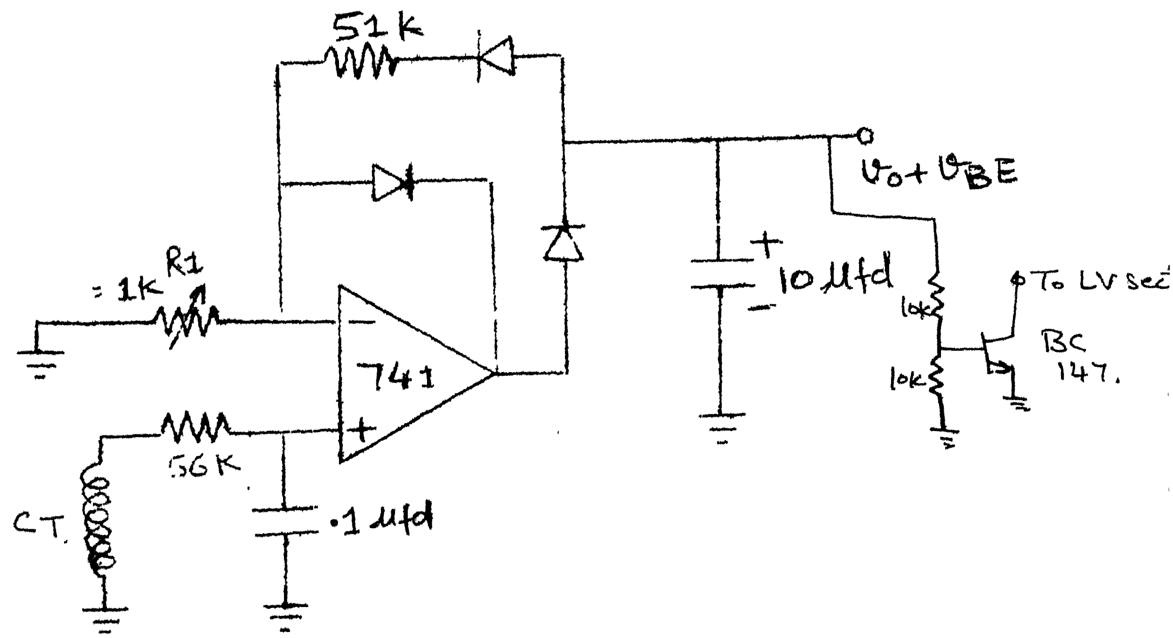


Fig 38.

10 μfd capacitor holds the rectified voltage. An IDMT (Inverse definite minimum time) stage is used to trip the relay. The IDMT is realized using a transistor integrator stage. To avoid tripping below 110% load a

reverse bias of .6 volts is applied so that the current source is only operative when input voltage is $>$ 0.6 volts. The IDMT works as following.

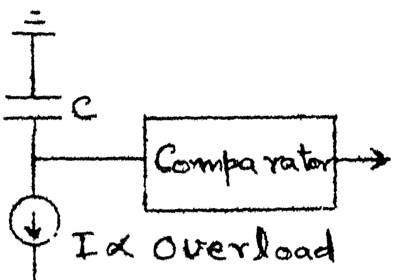


Fig 39.

Suppose we generate a current I α overload and let it charge a capacitor C, then $C\frac{dv}{dt} = I$ so $\Delta t = C\frac{V}{I}$

If we start from '0' voltage and switch a camparator at a fixed voltage V_f then

$$t = C\frac{V_f}{I}$$

so time α $\frac{1}{I\ \text{overload}}$ hence we get IDMT characteristics. The scheme is realized in the following way. as shown in fig 40.

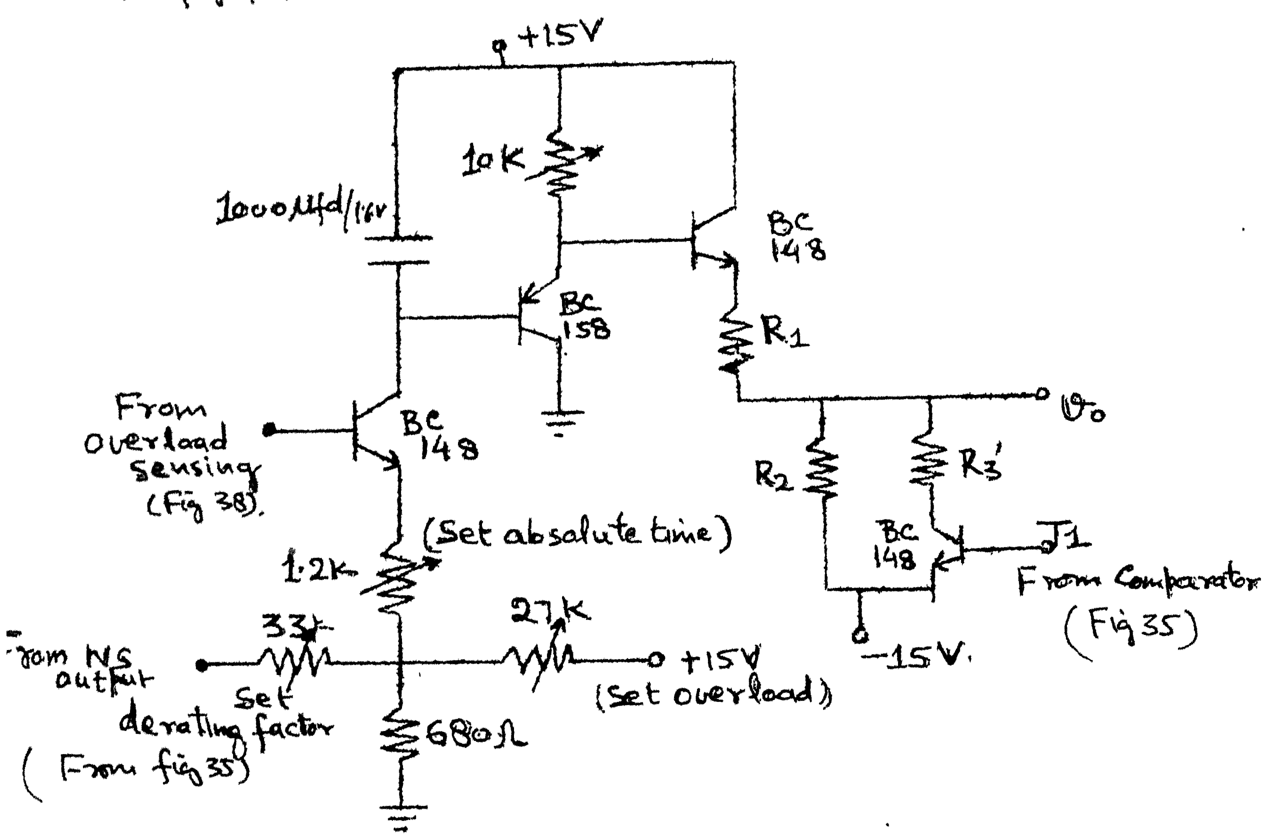


Fig 40

For 10 times overload v_i = 5 volts so at the collector of BC 148, we can have + 6 volts i.e. signal swing of (15-6) = 9 volts. At 6 volts we want v_o = − 3 volts so

$$\frac{6R_2}{R_1+R_2} - \frac{15R_1}{R_1+R_2} = -3.6 \text{ volts, say } R_1+R_2 = 30K$$

so $6R_2 - 15R_1 = -108$

or, $6(30 - R_1) - 15R_1 = -108$ or $-21R_1 = -288$

$$R_1 = 13K$$
$$R_2 = 17K$$

T_1 is normally off. When comparator switches to positive level, T_1 become on, so R_2 now has R_3 in parrallel, so the level at which comparator will switch again increases because the level of input is reduced. Thus we obtain indirect hysteresis for IDMT, while main comparator has only 100 mv. hysteresis. Suppose we want that canparater should switch back when 1000 μfd capacitor gets charged to + 14 volts. So,

$$\frac{14R_3}{R_1+R_3} - \frac{15R_1}{R_1+R_3} = -3 \quad (\text{where } R_3 = R_2 // R_3')$$

Therefore, $14R_3 - 15 \times 13 = -3(R_1+13)$

$$= -3R_1 - 39$$

$$17R_3 = 195 - 39 = 156$$

$$R_3 = 9K$$

So we chose R_3 = 10K

So the final circuit is as follows in fig 41.

$R_3' = 17K \| 17K \simeq 9K$

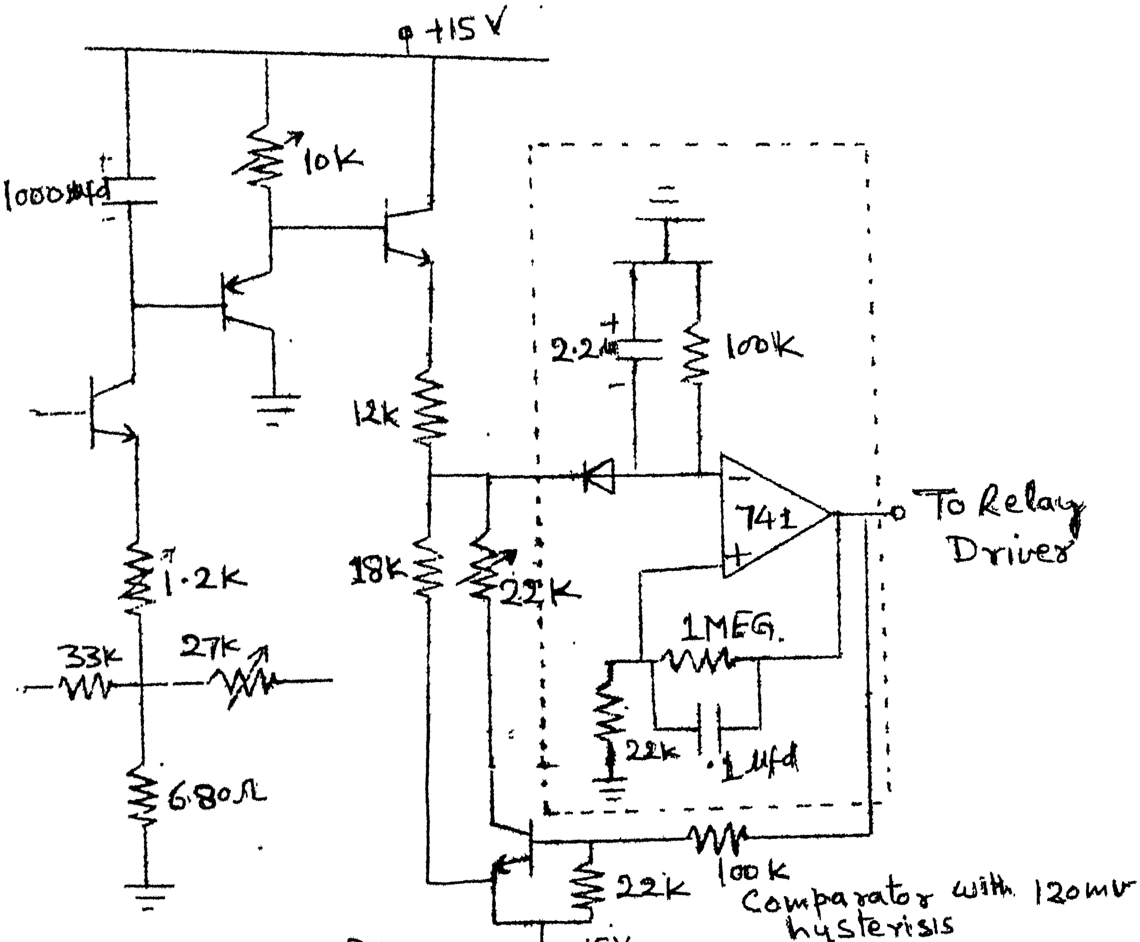


Fig 41.

For 100% overload i.e. v_1 = 1 volt

$$I = \frac{1-.5}{1.8} = \frac{.5}{1.8} = .3\text{m Amp.}$$

$$t = \frac{C\,V}{I} = \frac{1000 \times 10^{-6} \times 9}{.3 \times 10^{-3}} = 30 \text{ seconds}$$

For 200% overload

t = 15 seconds.

For 1000% overload

$t = 3$ seconds

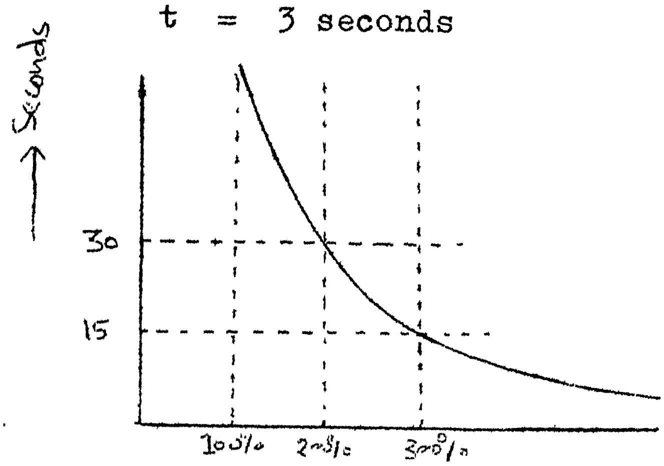


Fig 42.

To off time is determined by 1000 µfd and $\beta_{158} \times 10 \times 10^3$

$$\tau_{\text{discharge}} = 1000 \times 10^{-6} \times \beta_{158} \times 10 \times 10^3$$
$$= 10 \times \beta_{158} \quad 10 \times 200 = 2000 \text{ seconds}$$

$$v_{CO}\, e^{-t/\tau} = v_f$$

Therefore $9e^{-t/2000} = 1$

Therefore $e^{-t/2000} = \frac{1}{9}$

Therefore $1 - \frac{t}{2000} = \frac{1}{9}$ or $\frac{t}{2000} = \frac{8}{9}$

$t = 8 \times 220$ seconds

$\simeq 32$ minutes.

So motor gets 32 minutes as cooling time and can not be restarted manually if it trips on overload before this much time elapses. This time can be varied by varying 10 K pot.

If too many short duration overload occur then 1000 µfd capaciter acts as a memory and trips after many such short duration overloads. This feature of the relay is very much desired especially for large motors to avoid damage to imsulation by providing sufficient cooling to the motor.

To block low voltage operation when there is overload on motor the following ckt is used.

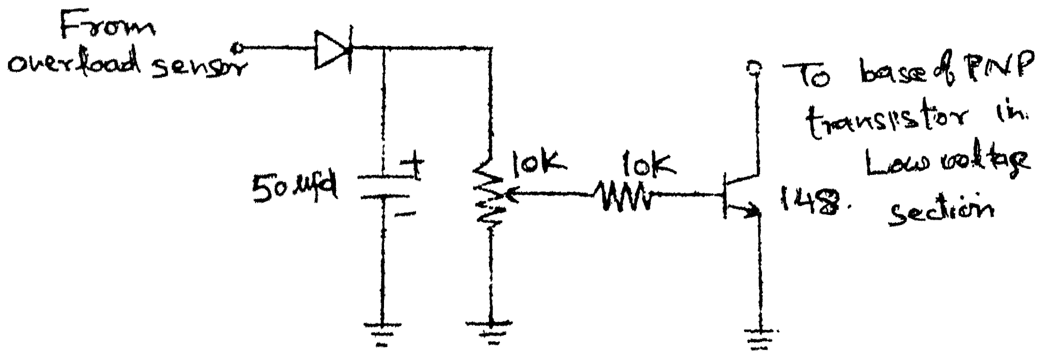
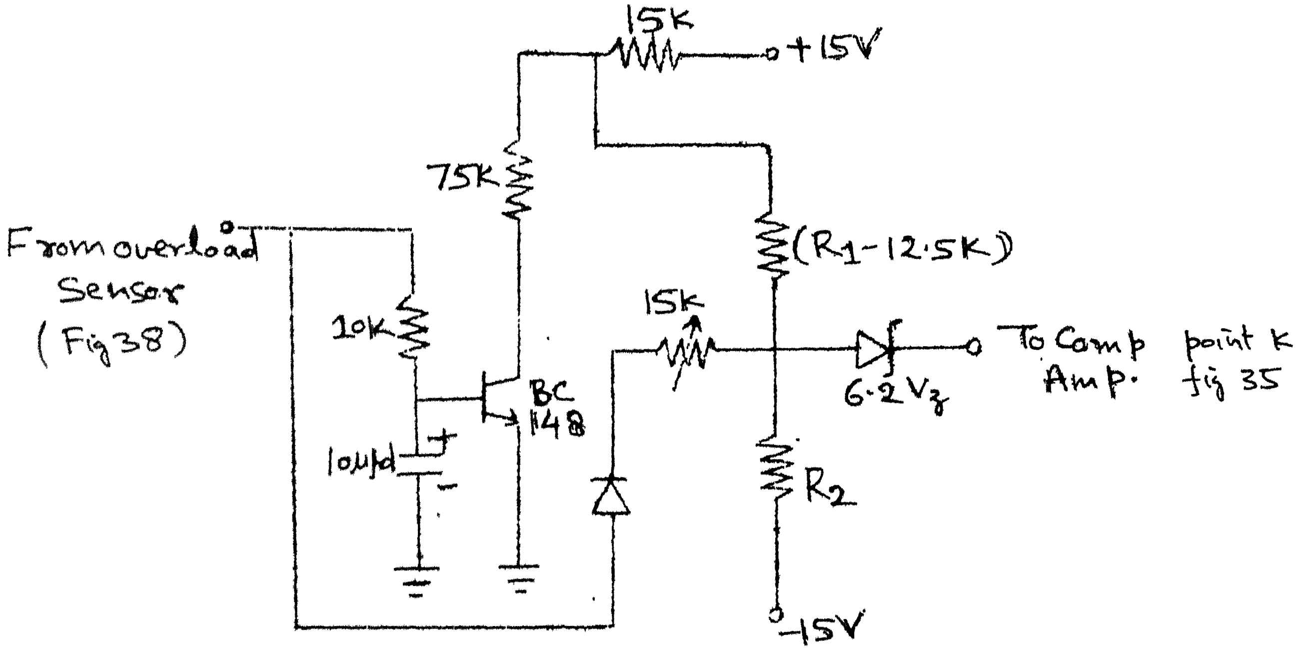


Fig 43.

The pot is adjusted so that when overload is about 20% the transister starts conducting and blocks low voltage operation.

$$\vartheta_o = 2\vartheta_d + (\overset{100\%}{.5} + \overset{20\%}{.1}) = 2\vartheta_d + .6 \text{ volts}$$

$$\vartheta_o' = 2\vartheta_d + .6 - \vartheta_d = \vartheta_d + .6$$



Low Load Protection Fig 44.

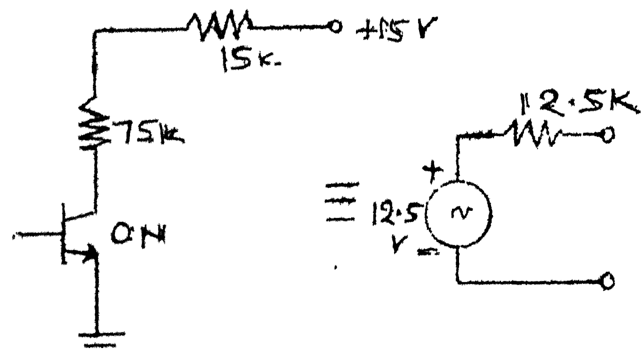


Fig 45.

$$\frac{12.5\,\frac{R_2 \times 15}{R_2 + 15}}{R_1 + \frac{15R_2}{R_2 + 15}} - 15 \times \frac{\frac{15\,R_1}{15+R_1}}{R_2 + \frac{15R_1}{R_1+15}} = -\ 6.2 \text{ volts}$$

$$\frac{12.5 \times 15\, R_2}{R_1R_2 + 15(R_1+R_2)} - 15 \times \frac{15\, R_1}{R_1R_2+15(R_1+R_2)}$$

$= -\ 6.2$

say $R_1 + R_2 = 56$ K

Therefore, $\frac{187.5R_2}{R_1R_2 + 840} - \frac{225R_1}{R_1R_2 + 840} = -\ 6.2$

$$-\ 31R_2 + 36R_1 = R_1R_2 + 840 = 56R_1 - R_1^{\ 2} + 840$$

$$R_1^{\ 2} - 20R_1 - 840 - 31\ R_2 = 0$$

or $R_1^{\ 2} - 20R_1 - 1680 + 31\ R_1 = 0$

or $R_1^{\ 2} + 11R_1 - 1680 = 0$

or $R_1 = -\frac{11 \pm \sqrt{121+6720}}{2}$

$= -\frac{11 \pm \sqrt{6840}}{2}$

$R_1 = \frac{-\ 11 + 83}{2} = \frac{62}{2} = 31$ K

Since 12.5 K is already there so $(R_1 - 12.5) \simeq 19$K

$R_2 = 56 - 31 = 25$ K

In fig 44. Transistor T_1 switches only when there is any finite load so this circuit is activated only when there is finite load otherwise while starting the motor it does not cause false tripping due to zero load condition. If after some time the load is not sufficient this section causes the motor to trip.

Underfrequency And Overfrequency Tripping :

Since the circuit compensates the N.S. voltage whenever there is a frequency drift in the supply, a simple technique is used to make the gain of main amplifier non linear by means of two zeners of $V_z = 2.0$ volts as shown in main amplifier. Underfrequency and overfrequency settings can be independently adjusted by means of the potentiometers in the circuit. This gives a continuously variable setting for frequency and may be very vital for operation during peak load season and low load season, especially where pumps or Ball mills are used.

Section Taking Care Of Long Accelaration Time, High On/Off Duty Cycle And Rapid Reversing :

Here one differentiater is used to find out the rate at which the current is decaying in the motor during accelaration. If this decayrate is more or equal to set rate as specified by the user then overload section is disabled from operation. But if this rate is less than the set rate the overload section is active and trips the motor depending on the amount of overload.

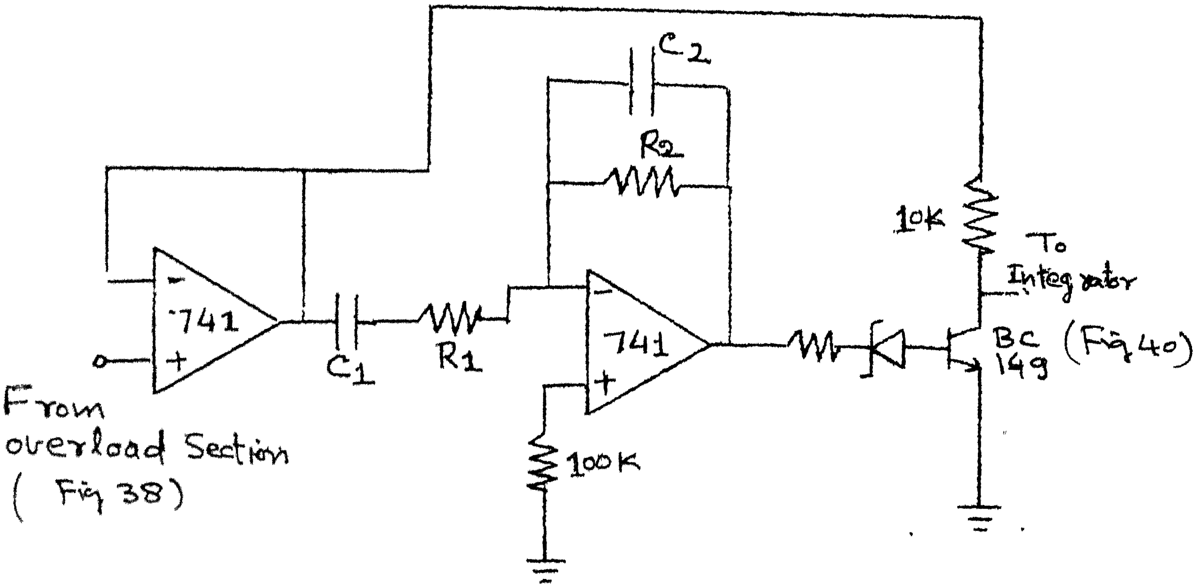


Figure 46.

$$R_2C_2 = \frac{1}{\omega} \quad \text{(Say 2 cycles)}$$

$$R_2C_2 = \frac{1}{12}$$

$$C_2 = 1\ \mu\text{fd polyster or Discceramic}$$

Therefore, $R_2 = \frac{10^6}{12} = \frac{100 \times 10^3}{1.2} \simeq 80K$

So for frequences above 20 cycles the capacitor C_2 acts as a short.

Since we have normalized the input voltage to .5 V = 100% load so we can set the rate of fall of voltage proportional to rate of fall of load.

Say the current becomes 10 times i.e. 1000% and it falls at a rate of 50%/sec. i.e. the voltage falls by - .25V/Sec.

$$\text{So } I = -C_1\frac{dv}{dt} = C_1 \times .25$$

Therefore, $\vartheta_o = -R_2 \times I = +R_2 \times C_1 \times .25$ suppose R_2 $\simeq 160K$

$$\text{So, } \vartheta_o = +160 \times 10^3 \times C_1 \times 10^{-6} \times .25$$
$$= +40 \times C_1 \times 10^{-3} \text{ volts}$$

C_1 in μfd is say $C_1 = 330\mu fd/15V$

Therefore, $\vartheta_o \cong +12$ volts.

BC 148 switches and short circuits the input of integrator to ground thus preventing its operation.

Now to prevent excessive currents for 50 cycle ripple and other transients from flowing through C_2 a resistor R_1 is introduced. This acts as a very low impedance compared to the impedance of capacitor at the desired frequency causing very little voltage drop across it but at higher frequencies it limits the input current.

$$I = C\frac{dv}{dt} = 330 \times 10^{-6} \times .25 = 80\ \mu\text{ amp.}$$

Voltage drop across capacitor = .25 volts.

For this resistor to act as a short circuit the drop across it should be .025 volts.

Therefore, $.025 = 80 \times 10^{-6} \times R$

Therefore, $R = 330\ \Omega$

So the final circuit becomes as following.

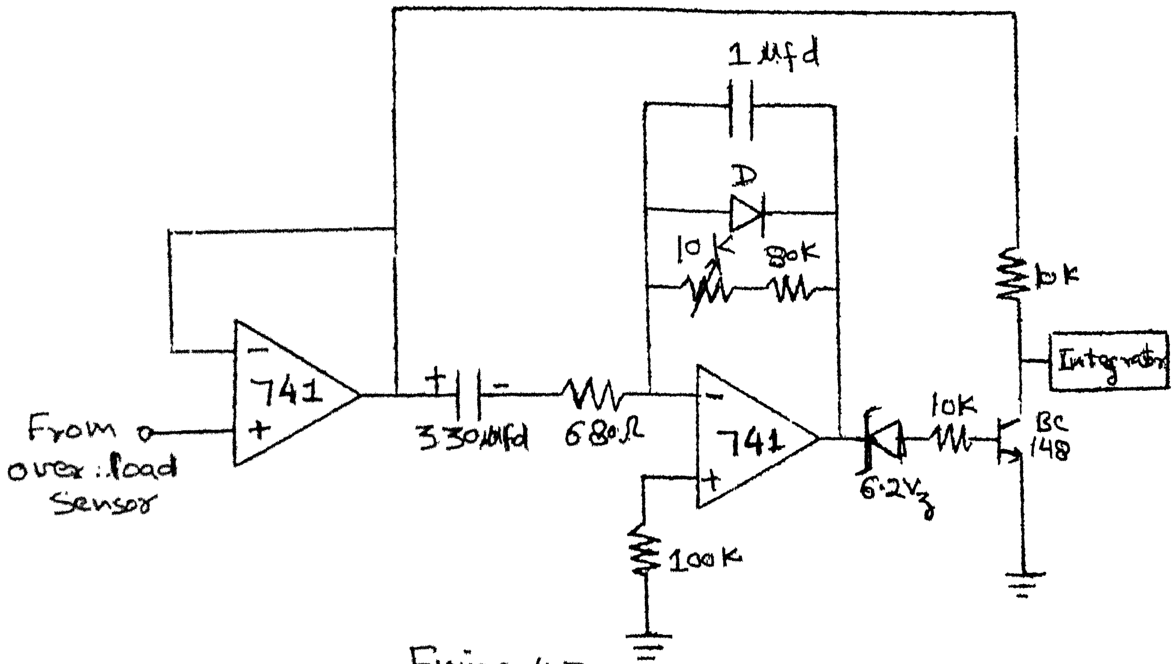


Figure 47.

Diode 'D' in the forward path is to avoid negative saturation of the OP Amp. when the current suddenly increase.

Earth Fault Sensing :

A separate Band pass amplifier is used to extract zero sequence signal from the motor. This is necessary to filter out third harmonic compenents from the current.These third harmonies might be generated either due to the saturation characteristics of the magnetic material of motor or it might arise due to the feeding transformer voltage distortion causing third harmonics to flow through the star grounded winding of the motor. A large number of secondary turns are wound over the CT to get a large output. This is then fed to a band pass amplifier and rectified to get DC

and fed to comparator. The operating current is 100 m Amp. The VC VS configuration is used for realizing band pass filter.

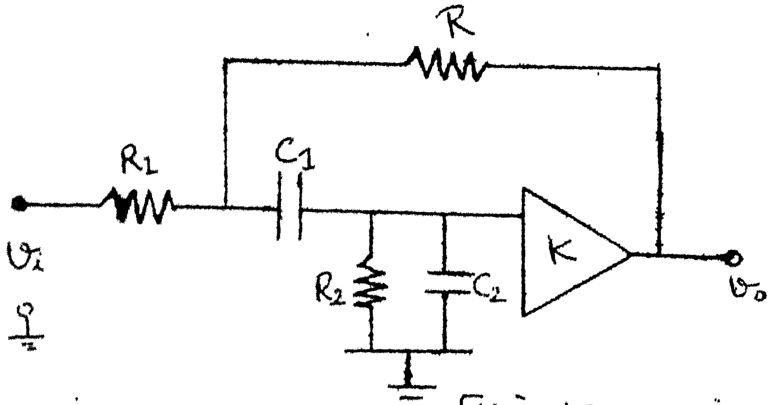


Figure 48.

$$\omega = \sqrt{\frac{2}{R_1 R_2 C_1 C_2}}$$ chosing $R_1 = R_2 = R$ and $C_1 = C_2 = C$

$$\omega = \frac{\sqrt{2}}{RC}$$

chosing $C = .05\ \mu fd$

$$R = \frac{\sqrt{2}}{\omega \times C} = \frac{1.414}{2\pi \times 50 \times .05 \times 10^{-6}} \simeq 82K$$

$$K = \frac{5 - \sqrt{2}}{Q}$$

For band width of 10 cycles. $Q = \frac{50}{10} = 5$

$K = 5 - \frac{1.414}{5} = 4.7$, so the actual circuit is as following.

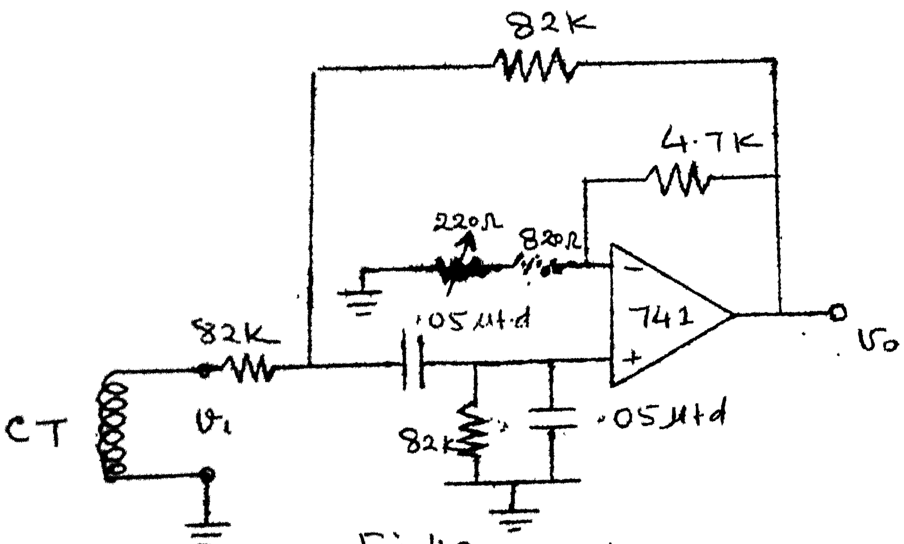


Fig 49.

CHAPTER VI

Design of a Three Phase Oscilator

Since the three phase supply available from distribution centres is generally unbalanced (magnitued and angle wise), for proper calibration of relays it is necessary to have controlled three phase supply. Since only two voltages are required, we need to generate only two voltage vectors and the third one can be generated by adding these two vectors to form a closed trangle. Since for deriving negative sequence quantities the difference form of vectors is used i.e. V_{ac} and V_{bc} etc., under balanced condition these two vectors have a phase shift of only 60°. Hence we need to generate two voltages which are 60° apart at the line frequency (50 Hz).

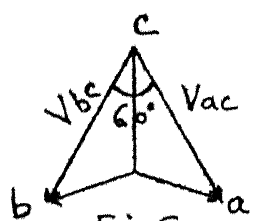
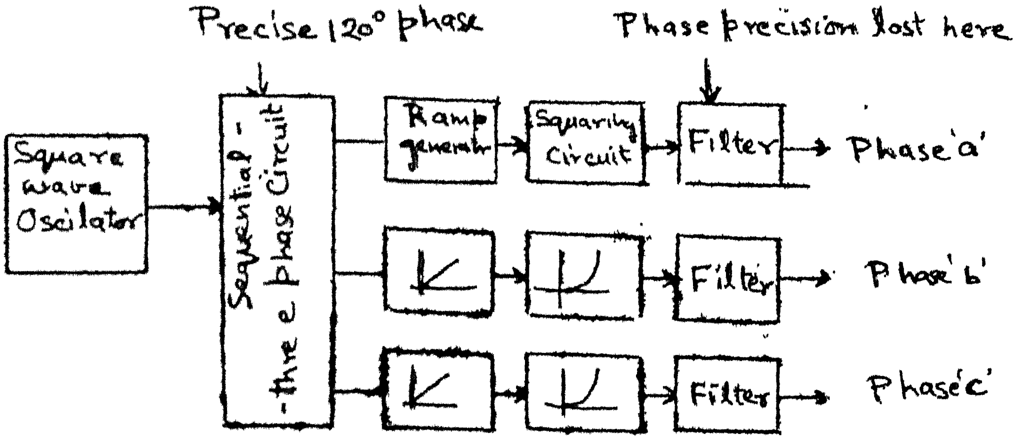


Fig 50.

Besides this to simulate actual generation condition the oscilater should have facility to vary the frequency over say 55 Hz to 45 Hz without variation in the phase difference between these two vectors and preserving their amplitudes.

For generating these vectors digital techniques may be used along with proper wave shaping and filtering, with the advantage that phase shift can be precisely maintained at 60°, but an overall review of these techniques reveals that



Technique 1 for generating three phase voltage
Fig 51

they are cumbersone and precision in phase obtained by digital circuit is ultimately lost by following analog circuits.

In fig 51. Another purely digital technique could be to use 8 bit- ROM programmed to contain samples of SIN wave in conjunction with Multiplexer and Demultiplexer and high frequency oscilator and could be addressed properly to maintain precise 60° phase shift. followed by D/A converter & filter.

Instead a very simple technique is utilized to get 60° phase shift by modifying the Wein-bridge oscilator a little. The principle is as following : Shown in figure 52.

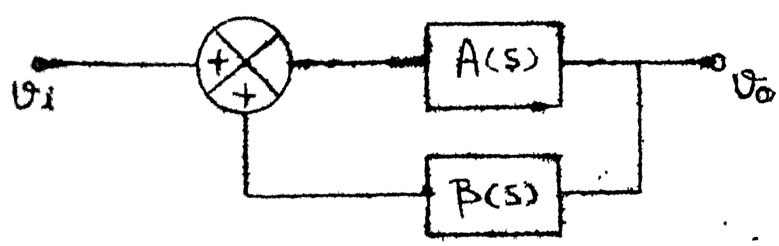


Figure-52.

For the above countrol system

$$\frac{v_o}{v_i}(s) = \frac{A(s)}{1 - A(s)\ \beta(s)}$$

For making an oscilator $1 - A(s)\,\beta(s) = 0$

or $A(s)\,\beta(s) = +1$ at desired frequency

If $A(s) = A$ i.e. frequency independent, then

$A\,\beta(s) = +1$ i.e. phase shift of the feed back network should be zero and loop gain should be unity at the frequency of oscilation. Now if we want to change the frequency of oscilation then we have to modify $\beta(s)$ now such that at this new frequency the above mentioned conditions are met. Now the advantage is taken of the fact that whatsoever may be the method of variation of $\beta(s)$ if the circuit is oscilating there will be zero phase shift around the loop and if carefully designed the phase shift at each node of $\beta(s)$ can be preserved irrespective of the frequency.

If instead of positive feed back we had chosen the negative feed back then

$$\frac{v_o}{v_i} = \frac{A}{1 + A\,\beta(s)}$$

For oscilation $A\,\beta(s) = -1$

Then loop phase shift required would have been 180°. To achieve this if we use three 60° phase shifters as in fig 53.

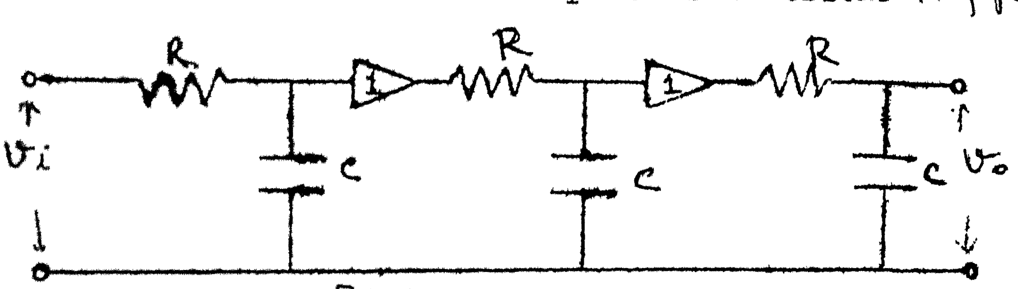


Fig 53.

In this case to vary the frequency and to preserve phase, all three resistors require simultaneous variation which is more cumbersome.

So if we resort to positive feed back and try to achieve '0' phase shift it is an easier task than achieving 180^0 phase shift. Say we modify the Wein bridge circuit as follows. in fig 54.

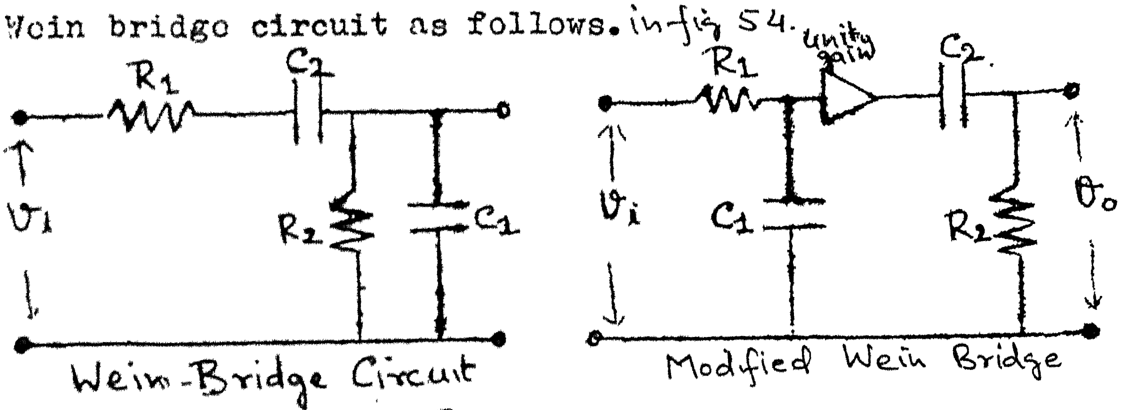


Fig 54.

The modified Wein-bridge acts as a low pass stage followed by a high pass stage. First stage causes a phase lag while second stage causes a phase lead and at some frequency the phase lag is compensated by phase lead to give overall zero phase shift between input and output. If now both R_1 and R_2 are increased in the same proportion the frequency at which the phase cancels out shifts by the same proportion.

Let us show it by rigorous analysis. For modified ckt,

$$\frac{v_o}{v_1} = \frac{1}{1 + SC_1 R_1} \times \frac{SC_2 R_2}{1 + SC_2 R_2}$$

$$= \frac{SC_2 R_2}{(1 + S^2 R_1 C_1 R_2 C_2) + S (R_1 C_1 + R_2 C_2)}$$

For phase shift to be zero.

$$1 + S^2 \; R_1C_1R_2C_2 \quad = \quad 0$$

or $$1 - W^2 \; R_1C_1R_2C_2 \quad = \quad 0 \text{ or } w = \frac{1}{\sqrt{R_1R_2C_1C_2}}$$

At this frequency

$$\frac{\vartheta_o}{\vartheta_i} = \frac{C_2R_2}{R_1C_1 + R_2C_2}$$

If R_1 and R_2 are increased by same factor, say K then

$$w_K = \frac{1}{K \; R_1R_2C_1C_2} = \frac{W}{K}$$

Phase shift at w_K in phase lag network

$$\theta w_{K_1} = \tan^{-1} w_K \, KR_1C_1 = \tan^{-1} \frac{W}{K} K \; R_1C_1 = \tan^{-1} wR_1C_1$$
$$= \theta w_1$$

i.e. phase shift is invariant of the change but only restriction being that R_1 and R_2 has to change by same factor. Since ganged potentiometers are available we may chose $R_1 = R_2 = R$

So,

$$\frac{\vartheta_o}{\vartheta_i} = \frac{SC_2 \; R}{(1 + S^2R^2C_1C_2) + S\,(C_1+C_2)\,R}$$

Frequency of oscilation $= W = \frac{1}{R\sqrt{C_1C_2}}$

at this frequency gain

$$= \frac{C_2}{C_1 + C_2}$$

T.F of first network $= \frac{1}{1 + j\,w\,RC_1} = \frac{1}{1+j\,\frac{1}{R\sqrt{C_1C_2}}\,RC_1}$

$$\text{T.F}_1 = \frac{1}{1 + j\,\frac{\sqrt{C_1}}{C_2}} \qquad \dots. \quad (14)$$

$$\phi = \tan^{-1} \sqrt{\frac{C_1}{C_2}}$$ .... (15) which is independent of frequency of oscilation

Similarly $TF_2 = \frac{j\,W\,R_2C_2}{1 + j\,W\,R_2C_2}$ substituting $w = \frac{1}{R\sqrt{C_1C_2}}$

$$= \frac{j\sqrt{\frac{C_2}{C_1}}}{1 + j\sqrt{\frac{C_2}{C_1}}}$$ this is also independent of frequency of oscilation

Since we are interested in a phase shift of 60^o going back to equ. 15

$$\tan 60 = \sqrt{\frac{C_1}{C_2}} = \sqrt{3} \quad \text{Therefore, } C_1 = 3C_2$$

gain of first network $= \frac{1}{1 + j\sqrt{3}} = \frac{1}{2}$

gain of second network $= \frac{j\frac{1}{\sqrt{3}}}{1 + j\frac{1}{\sqrt{3}}} = \frac{j}{j + \sqrt{3}} = \frac{1}{2}$

overall gain with buffer $= \frac{1}{2} \times \frac{1}{2} = \frac{1}{4}$

Since the frequency is to be varied there is likelyhood that resisters R_1 and R_2 may have some mismatch and may not give correct phase shift so we revert back to orignial equations.

$$W = \frac{1}{\sqrt{R_1R_2C_1C_2}}$$

$$TF\,1 = \frac{1}{1 + j\,\frac{\frac{1}{R_1C_1}}{\sqrt{R_1C_1R_2C_2}}}$$

$$= \frac{1}{1 + j\sqrt{\frac{R_1C_1}{R_2C_2}}} = \frac{1}{1 + j\sqrt{\frac{R_1}{R_2}}\sqrt{\frac{C_1}{C_2}}}$$

$$= \frac{1}{1 + j\sqrt{3}\sqrt{\frac{R_1}{R_2}}}$$, Capacitor ratio can be fixed precisely once for [illegible]

Since we want $\pm$ 10% frequency variation both the resistance should change by $\pm$ 10% around normal value. Let us consider the worst case where R_1 and R_2 are off by 10% (in actual case even with 10% tolerance resistances they will not be off by more than 2%) then

$$\frac{R_1}{R_2} = 1.1, \text{ say}$$

then $$TF\ 1 = \frac{1}{1 + j\sqrt{3}\sqrt{1.1}} = \frac{1}{1 + j \times \sqrt{3} \times 1.048}$$

$$= \frac{1}{1 + j\ 1.81}$$

$$\phi = \tan^{-1} 1.81 = 61^{\circ} \text{ (one degree more than normal)}$$

This surprisingly low sensitivity to phase shift is due to the fact that tan θ V_s θ rises very steeply above 45° and hence even a large change in the imaginery term causes very little variation in the phase angle as shown below .

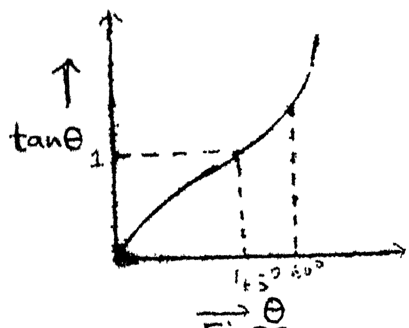


Fig 55

gain of amplifier for oscilation is

$$A = \frac{R_1C_1 + R_2C_2}{R_2C_2}$$

$$= 1 + \frac{R_1C_1}{R_2C_2}$$

$$= 1 + 3\frac{R_1}{R_2}$$

$$= 1 + 3 \times 1.1 = 4.3$$

So this requires non linear amplitude stabilization technique for selt adjustment of gain under such mismatched conditions.

The overall scheme is realized using three opamps as shown in figure II. OP Amp 1 is the *FET1 controlled by OP Amp 3 is used to get non linear gain control and thus providing stable amplitude. OP Amp 2 is used as a buffer as well as amplifier with gain 2 with low output impedence for phase 2. FET 2 is matched with FET 1 to compensate for amplitude variation in output of phase-II and to track with phase-I output. The control voltage for FET 2 is derived from FET 1 control voltage.

* forward Amp with gain 2

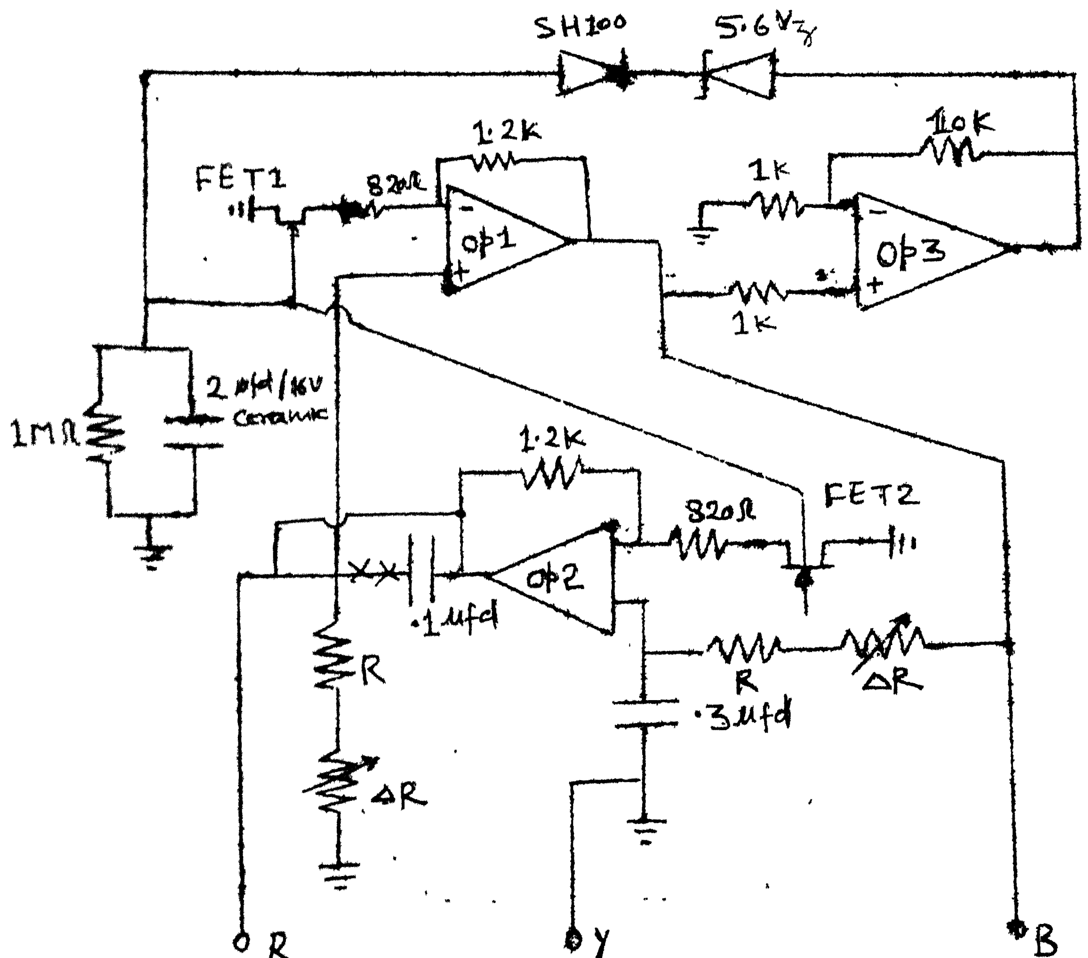


Fig 56.

The reason for using FET's in conjunction with each amplifier is as follows.

$$TF\ 1 = \frac{1}{1 + j\sqrt{3}\sqrt{\frac{R_1}{R_2}}}$$

$$TF\ 2 = \frac{j\sqrt{\frac{R_2}{R_1}}\ \frac{C_2}{C_1}}{1 + j\sqrt{\frac{R_2 C_2}{R_1 C_1}}} = \frac{1}{j + \sqrt{3}\sqrt{\frac{R_1}{R_2}}}$$

At centre fequency of 50Hz say $R_1 = R_2$

$$TF\ 1 = \frac{1}{\sqrt{1 + \frac{3R_1}{R_2}}} \quad \begin{matrix} 50Hz \\ R_1 = R_2 \end{matrix} \quad = 1/2 \quad \ldots\ldots \quad (15)$$

$$TF\ 2 = \frac{1}{\sqrt{1 + \frac{3R_1}{R_2}}} \quad 50Hz \quad = 1/2 \quad \ldots\ldots \quad (16)$$

As we vary R by varying ΔR, there may occur mismatch between R_1 and R_2 thus the gain of each R-C network changes but as is evident from equs (15) and (16) the gain is equal for both the networks at the frequency of oscilation. This fact has been utilized by employing FET's in both OP Amps 1 and 2. If only one FET in OP Amp had been employed total gain change to maintain oscilations under mismatched conditions would have occured in OP Amp 1 and only phase 1 output would have been stabilized. The gain of OP Amp 2 in that case would have been say 2, then phase II output at frequency other than 50Hz would have been

$$V_{phase_2} = \frac{V_{phase_1}}{\sqrt{1 + 3\frac{R_1}{R_2}}} \times 2$$

If FET 1 and FET 2 and feed back resistors in OP 1 and OP 2 are matched then gain of both the amplifiers will track and since the gain of each network at the frequency of oscilations tracks each other, the variation in gain of network 1 will be balanced out by variation in gain of OP Amp II due to presence of FET - II and vice versa for network 2. Overall gain of networks

$$= \frac{1}{\sqrt{1 + \frac{3R_1}{R_2}}} \times \frac{1}{\sqrt{1 + \frac{3R_1}{R_2}}} = \frac{1}{x^2} \text{ say}$$

Gain of amplifiers required = x^2

Since both the amplifer employ matched FET and same control voltage so gain of each amplifier due to negative feed back becomes = x.

Phase 1 output is stabilized due to nonlinear feed back

Therefore,

$$\text{Phase 2} = \frac{V_{ph1}}{TF\ 1} \times \text{Gain of OP Amp 2}$$

$$= \frac{V_{ph1}}{x} \times x = V_{ph1}$$

So both the outputs track each other.

Now to simulate different kind of imbalances in three phase supply, the outputs are taken to control amplifiers where these two outputs can be mixed in to get different phase and amplitude outputs when desired. The circuit diagram is shown in figure ~~III~~.57 on next page.

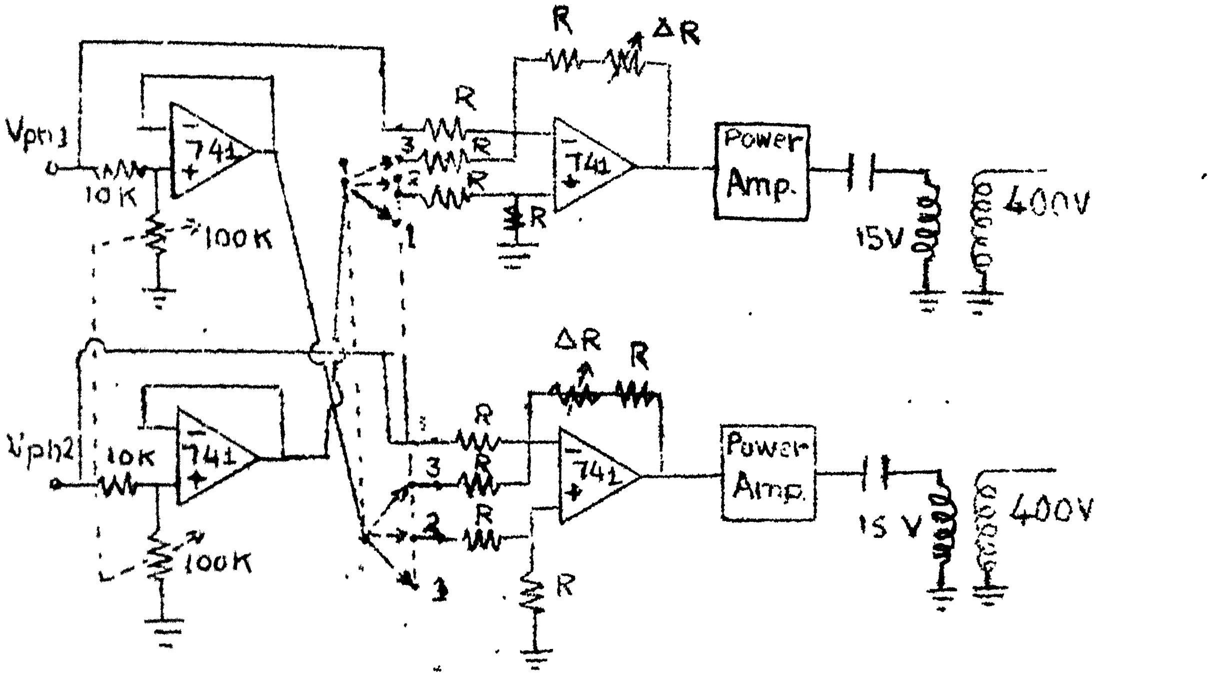
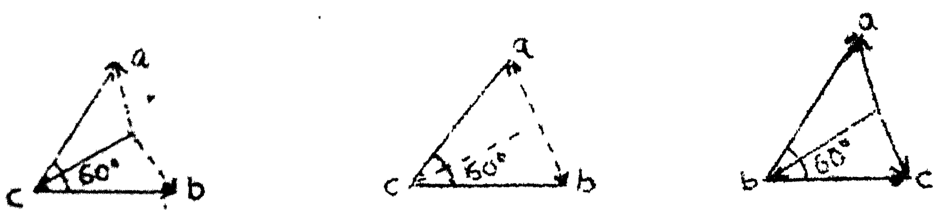


Figure ~~III~~ 57.

Model 1 : When switch is in position 1. Following vectors are available by varying R.



Model II : Switch in position 2. By varying look pot and R we get following vectors.

Model III : Switch in position 3. By varying look pot and R we get following vectors.

Power Amplifiers :

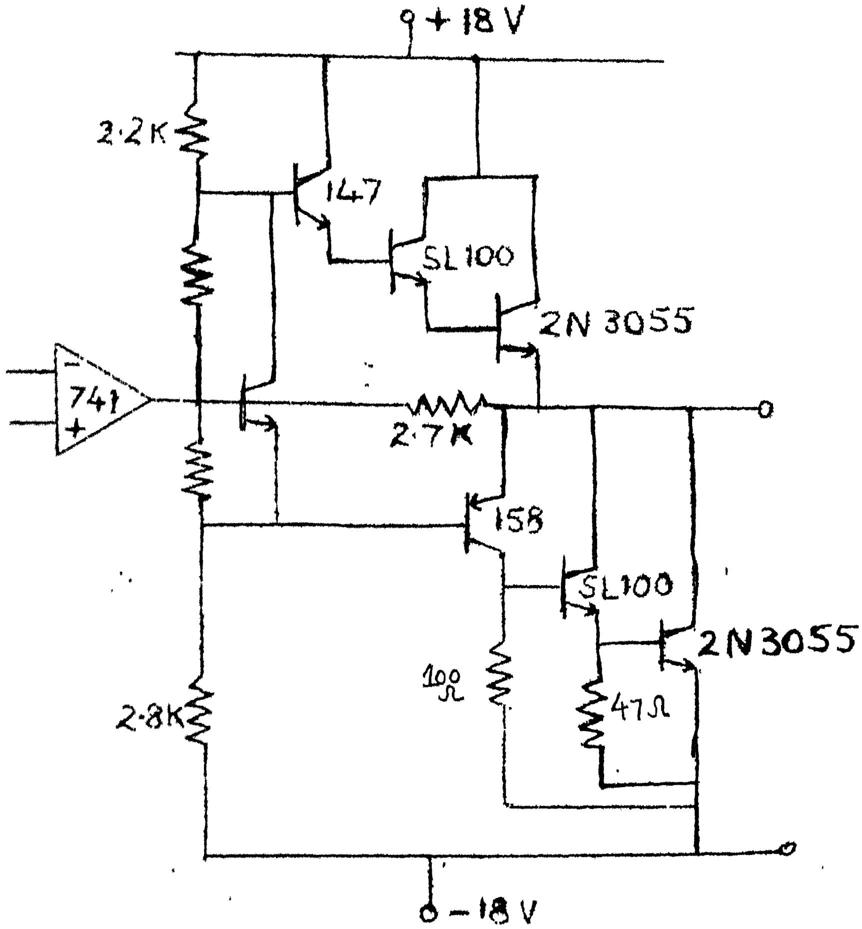
Two power amplifiers each of 55 Watts is used with each phase alongwith a step up transformer to boost the voltage to 440 Volts. OP Amp 741 is used as preamplifier since the frequency is only 50Hz.

The circuit diagram of the power amp is as shown in figure ~~IV~~ 58. The output is derived through a step up T/F 440 V/15 Volts having magnetizing current of approx. 10 m Amp. hence the secondary magnetzing current becomes.

$$I_2 = \frac{440}{15} \times 10 \text{ m Amp.} \quad 300 \text{ m Amp.} = .3 \text{ Amp.}$$

So this becomes the lagging current but in a power Amp this is dissipated as real power. At 440 V we can draw upto about 150 m Amp which is equivalent to 4.5 Amp. primary current.

The composite circuit alongwith the TIF is shown in figure ~~V~~. 59.



Power Amp 55 Watts

Figure ~~IV~~ 58.

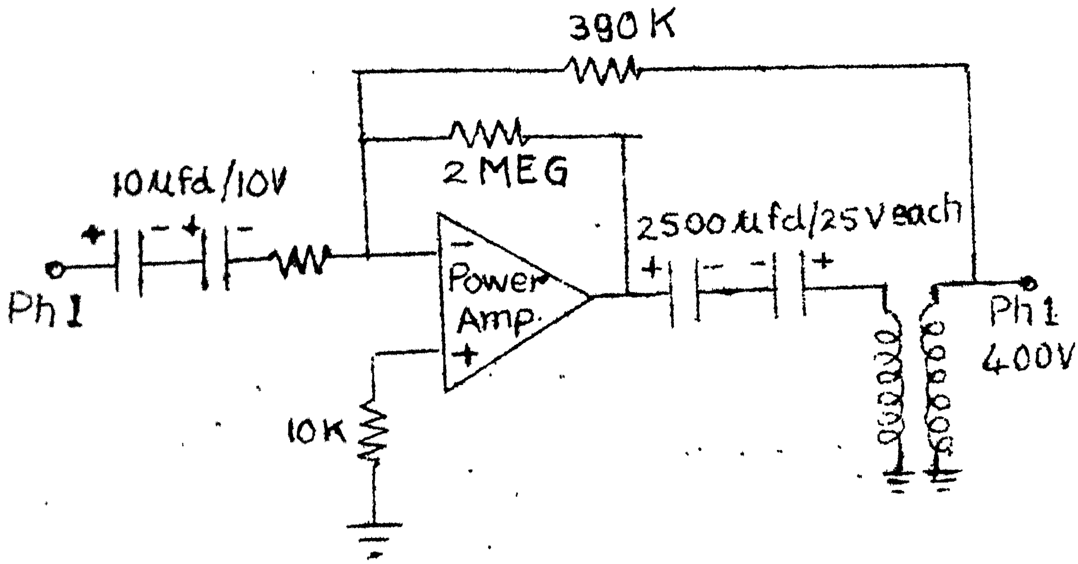


Figure #59

CHAPTER VII

Results and Conclusions

Different blocks of the unit were tested after assembling them as per the designs given in different chapters and satisfactory performance was obtained.

The analysis of induction motor done in Chapter II conforms to the practical results obtained from test on machines which are listed in the same chapter.

The frequency compensation circuit for Negative Sequence filter exhibits excellent compensation characteristics over frequency range from 45 Hz to 55 Hz. The % compensation vs. frequency is plotted in the following graph.

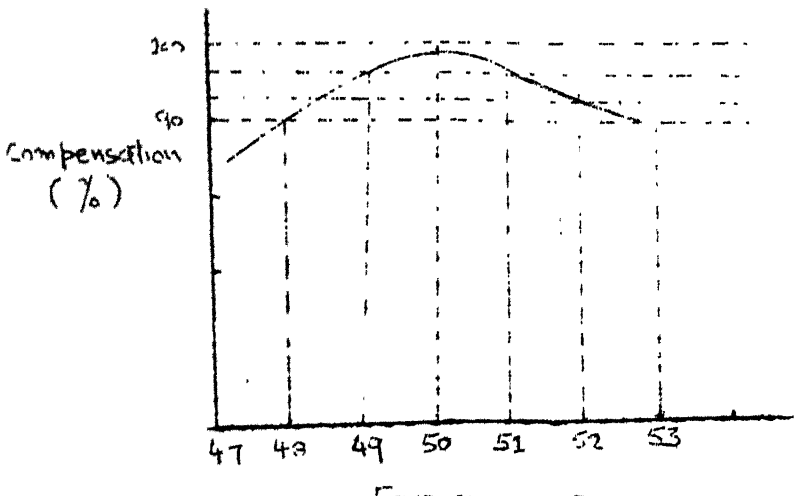
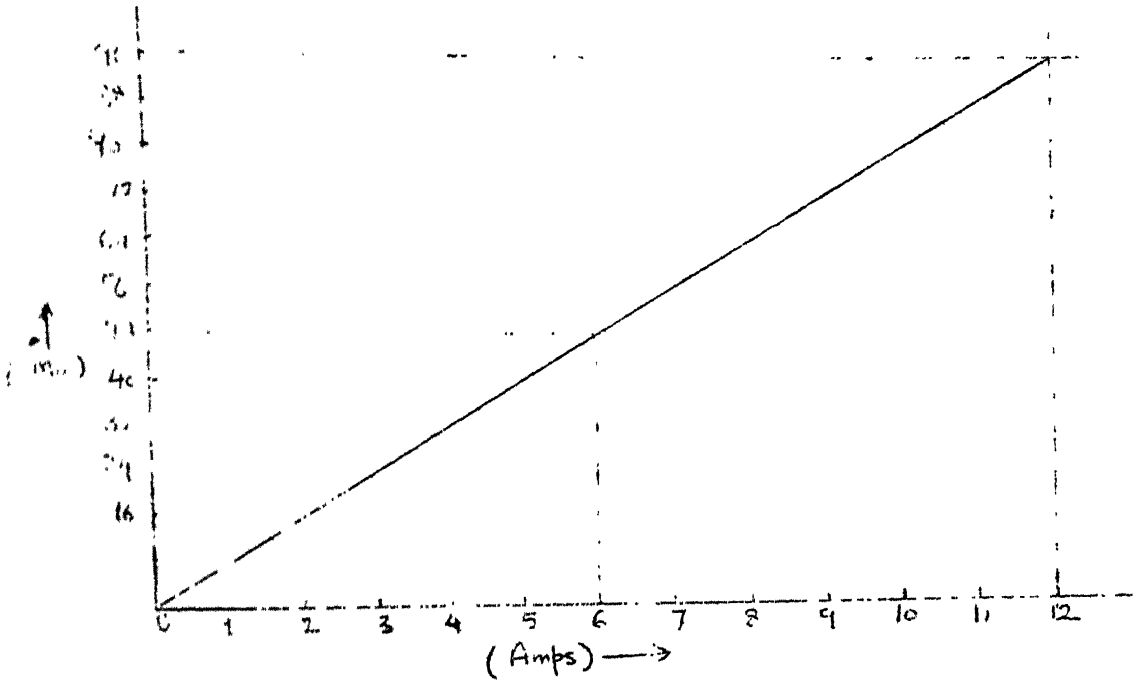


Fig 60.

These compensation characteristics can be improved by better matching of R & C values in Negative Sequence and compensation circuits but for notmal operating range they are good enough.

The CVT described in Chapter IV was wound over a bakelite toroid of 1¼" diameter with 500 turns of output coil and 5 turns of primary. It was tested upto 12 Amps. along with an R-C integrator. The sensitivity was 8 mV/Amp. The input output characteristic is absolutely linear.



CVT CHARACTERISTICS
Fig 61

With proper setting in the Negative sequence filter the unit trips at 5% unbalance in voltages at 50 Hz. The output of absolute value circuit to an applied sine wave is rectified S the wave. as shown in fig 62.

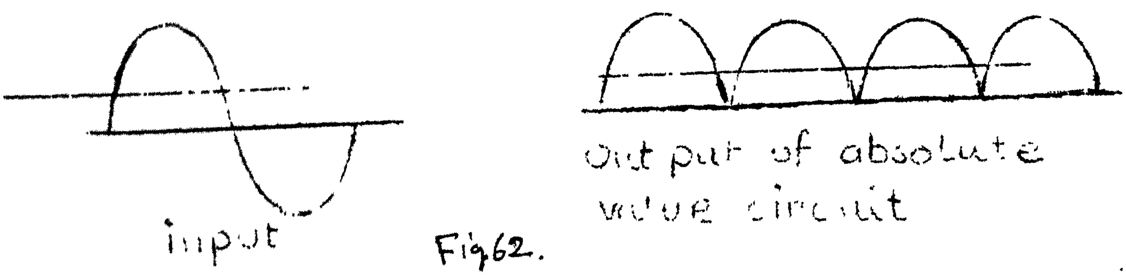


Fig 62.

This circuit alongwith compensation and N.S. voltage gives true negative sequence voltage at all frequencies of interest.

The tripping characteristics of the relay under different input voltage and input current conditions are given in the graph below. It exhibits a linear characteristic in the low voltage region with input current.

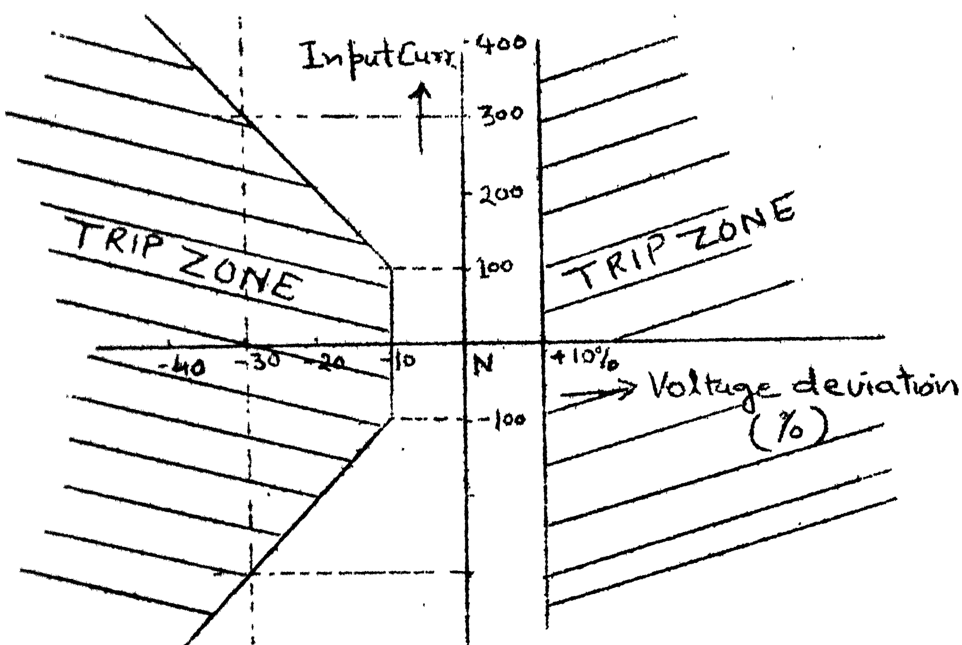


Fig 63.

Tripping characteristics, input voltage Vs. Input Current.

The derating feedback provides additional protection to motor against overheating under unbalanced conditions of voltage. The motor rating vs. unbalance in line voltage is shown in the following figure.

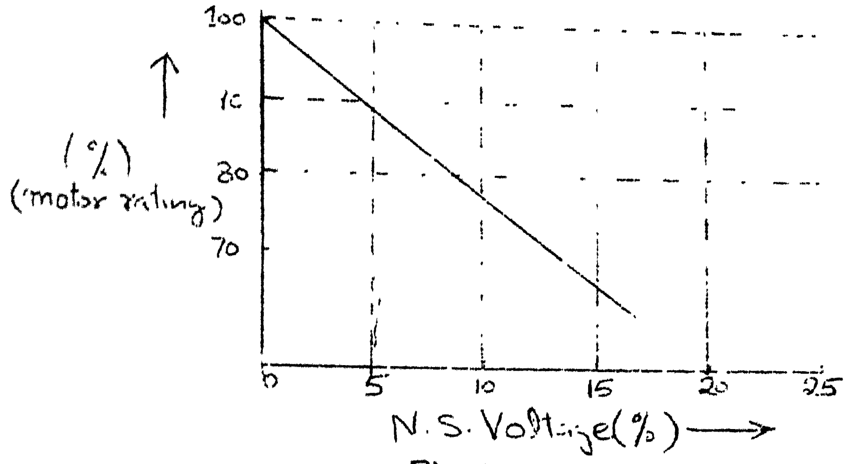


Fig 64.

The tripping characteristics of overcurrent vs. acceleration are shown in the following figure.

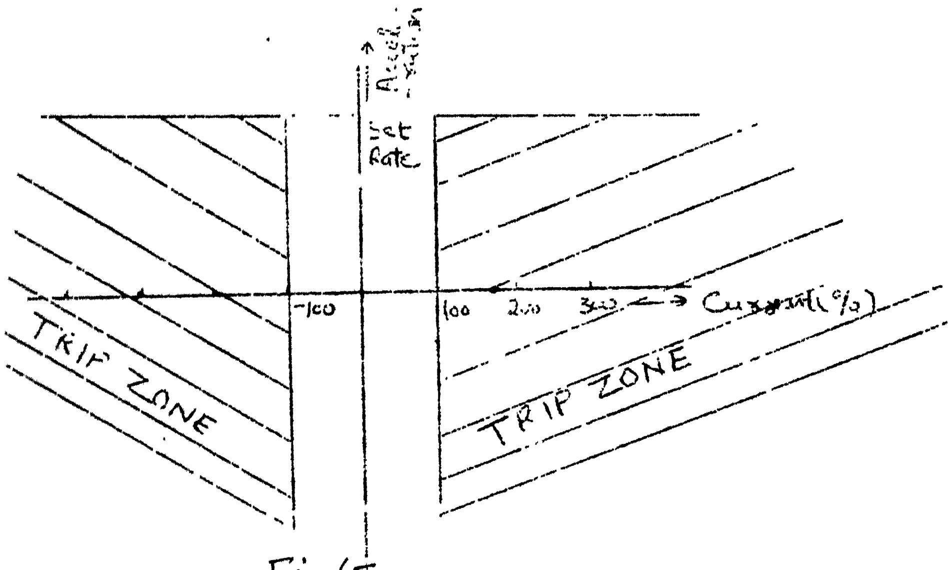


Fig 65.

Tripping characteristics, input current vs. Acceleration

The cooling time for medium range motors has been kept 5 minutes and this time does not alter even if the power supply fails during the time when motor is cooling off. Even if power is restored again the motor will not start till the set time has elapsed. This feature results in considerable reduction in power consumption of the unit because cooling timer does not require any energy from the source rather it dissipates the extra energy stored in capacitor due to overload in motor.

The three phase oscillator gives sine waves of equal amplitude at frequencies between 45 Hz to 55 Hz maintaining a phase difference between 60-60.5 degrees. This oscillator can be used as test power source for any power equipment by suitably modifying the power amplifier.

Scope for further work:

For larger motors above 70 h.P. it may be worth-while to evaluate the economic feasibility of a microprocessor based protection scheme incorporating all the above features. Besides this, the protection of large motors may require additional thermal protection to monitor actual states temperature. For motors between 50-70 hp the insulation tolerance limit may require the implementation of inverse I^2 characteristics which may be a little uneconomical and unnecessary for lower rating motors. The schematic diagram

for deriving $1/I^2$ characteristics is given below. in fig 66.

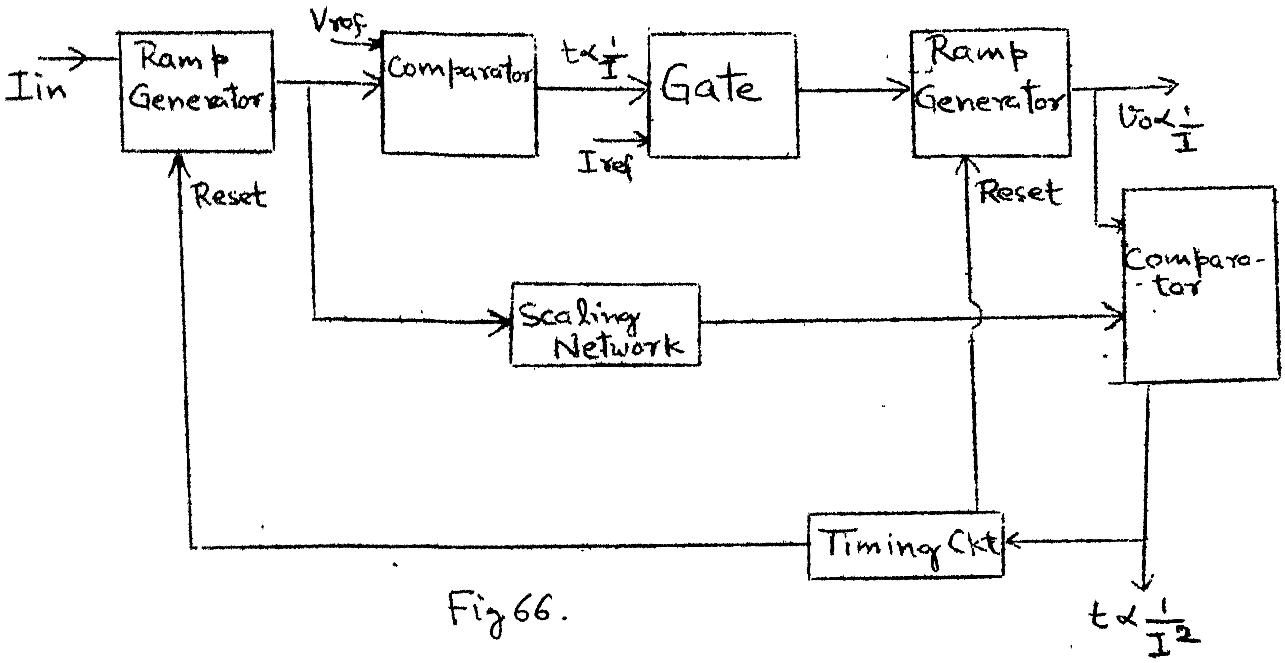


Fig 66.

Conclusion:

For motors ranging between 1.5 to 50 hp it is expected that the proposed scheme of protection is sufficient to cater for all normally encountered faults, reliably and repeatedly. It should prove to be a cheap, fast and efficient scheme for all type of industrial and commercial applications of induction motors. If properly tested and fabricated this is expected to avoid a large number of damages to motors which bimetallic relays are not able to take care due to their inherent limitations and overdesign.

The analysis of Induction motors shows that when one of the phase fuses blows, a Negative sequence voltage of the order of 5% is generated across the three phases at full speed. Hence the Negative sequence voltage sensing circuit should trip before this much, voltage disbalance occurs. In addition to this the motor should be dynamically derated when the voltages are unbalanced to avoid overheating in the roter and stator. For small machines their derating factor is 2.2% derating in current for 1% N.S. Voltage. Both these have been incorporated and tested satisfactorily in the present scheme.

The frequency dependence of the Negative sequence voltage sensing circuit has been nullified by using proper compensation circuit. This feature is very important for proper functioning of the whole scheme under unbalanced conditions as frequency does not stay at 50 Hz constant.

The heart of the voltage section is an absolute value cell using two Op-Amps. This cell performs all the voltage functions desired. This handles low voltage, high voltage, frequency compensation, N.S. Voltage and no load signals. This results in considerable saving of hardware thus reducing the overall cost.

In the current control section an IDMT stage has been used. This has got separate controls for setting absolute time, overload setting and derating factors. This also provides certain minimum cooling time to the motor once it

trips on overload. This cooling time is independent of power failure during this mode.

An air core CT with proper IC integrater has been used to generalize the design for a very wide range of motor ratings and at the same time reduce cost with improved linearity.

For proper caliberation of the relay a simple modified wein bridge has been utilized for generating three phase supply. The frequency of this can be varied between 45-55 Hz without effecting the relative phase and amplitude. Independent control of amplitude also provides the facility to generate unbalanced conditions.

REFERENCES

1. Vickers, Induction Motors, McGraw Hill Publication.

2. Evans and Wagner, Symmetrical Components, International Student Edition.

3. Alger, Theory of Induction Machines, John Wiley and Sons.

4. Dover, Theory and Practice of Alternating Currents.

5. Burr and Brown, Operational Amplifiers; Design and Applications, International Student Edition.

6. Thermal Tracking - A Rational Approach to Motor Protection by David R. Boothman, Everette Elgar, Trans. IEEE PAS Sept.-Oct., 1974.

7. Fast Response Detection of Mean Value of Power Systems Quantities by R. Ueda, E. Ohta, H. Nakashima, Trans. IEEE PAS-94, Dec. 1975.

8. Frequency Deviation Transducer for Power System Applications, Prabhat Kumar, Trans. IEEE PAS July-Aug. 1975.

9. Transient Response of Current Transformer - Committee Report, Trans. IEEE PAS Jan. 1976.

10. Inverse Time Overcurrent Relays using Linear Components by H.K. Verma, T.S.M. Rao; Trans. IEEE Vol. PAS-95, Sept.-Oct. 1976.

11. Current Transformer Accuracy with Asymmetric and High Frequency Fault Currents by D.A. Douglass, Trans. IEEE PAS Vol. 100, March 1981.

12. Power System Protection - Static Relays by Madhav Rao, Tata McGraw Hill.

13. Circuit Analysis of AC Power Systems, Vol. I and II by Edith Clarke, John Wiley.

14. Switchgear and Protection - S.S. Rao, Khanna Publishers.

15. The Art and Science of Protective Relaying by Mason, Wiley Eastern.

16. Optoelectronic Current T/F in Russian by A.Fanasev, Electrichestro, July 1970.

17. Instrument Transformers, Trans. IEEE, IGA-6, Nov.-Dec. 1970, Committee Report.

18. Transient Behaviour and Uses of Current Transformers, by N.Korpanay, Brown Boveri Review, Nov.-Dec. 1969.

19. Design of Current Transformers for Measurement of Heavy Currents, by S. Seely, Electrical Review, June 1970.

20. Theory and Design of Electronic Relays, by L.O.Chua, Proc. IEEE, Nov. 1970.

21. Linear Applications Notes - National Semiconductors, U.S.A.

22. Electric Machinery by A.E.Fitzgerald and Charles Kingsley, Jr., International Student Edition.

23. Pulse, Digital and Switching Waveforms, Millman and Taub, International Student Edition.

24. Electronic Circuits, Millman and Halkias, International Student Edition.